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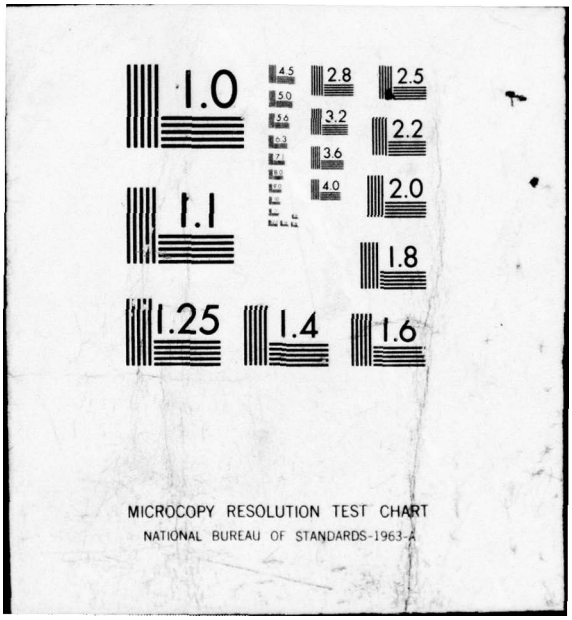
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## PROJECT FIST

(Fault Isolation by Semi-Automatic Techniques)

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(Fault Isolation by Semi-Automatic Techniques)

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## PREFACE

Fault Isolation by Semi-automatic Techniques is offered as an immediately applicable solution to the problem of maintaining new, non-computer electronic equipment. The techniques have been reduced to practice and find immediate application in the measurement of the most commonly-encountered circuit functions. The number and type of tests made on each piece of electronic equipment is completely within the control of the equipment designer, who can apply the basic transformation network techniques without requiring the intervention of the instrumentation or test equipment specialist.

It is felt that the FIST maintenance system should meet with ready acceptance because of its flexibility. The number of FIST-maintained electronic equipments that are used in an installation may change in number and variety without interfering with the use of FIST on any of the other equipments. The system can be used economically even if only a single prime system is built to use the technique.

Part I of this report introduces the FIST concept, explains the principles upon which it is based, examines techniques for physically incorporating FIST into the prime equipment, and describes the features which give the method its flexibility. Part II describes the transformation networks used to convert a variety of signal characteristics to voltages which can be measured using the FIST test instrument, and Part III describes a prototype test instrument. Part I is sufficiently complete to enable the reader to evaluate the usefulness of the FIST system without reference to Parts II and III. Parts II and III will be useful to the reader who is interested in examining the FIST concept in more detail.

Gustave Shapiro
Owen B. Laug
George J. Rogers
P. Michael Fulcomer, Jr.

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# PART I

## FIST GENERAL PRINCIPLES

### INTRODUCTION

Project FIST (Fault Isolation by Semi-Automatic Techniques) has been conducted at the National Bureau of Standards under the sponsorship of the Department of the Navy, Bureau of Ships. Application of these techniques is biased toward shipboard installations and other areas where small size and weight, though important, are not overwhelmingly so, and where a moderate increase of size and weight may be profitably traded off against a practical maintenance capability. The system is one that must be included in the initial design of the prime equipment if its full potential is to be realized.

When considering fault isolation techniques, it appears logical to divide all electronic devices into two groups: computer systems (both digital and analog), and non-computer systems. Perhaps the most efficient way to diagnose faults in computer equipment is to program test problems, along which lines much work has been done. However, in non-computer areas such a variety of circuits and functions are encountered that no general method (other than manual) has previously been demonstrated to have wide utility. For these reasons, a decision was made early in this program to concentrate on the development of those techniques that would best fit non-computer applications. Although these techniques could be used to isolate faults in computer equipment, it is suggested that other techniques specifically tailored to computer needs may be more efficiently employed.

This development is concerned with techniques that will permit untrained technicians operating under field conditions to make meaningful performance measurements on electronic assemblies after 5 or 10 minutes of simple instruction. It is intended that the untrained technician, with the aid of a portable general purpose test set, be able to go into a room full of equipment that had been designed to work with the FIST maintenance system, find and rapidly localize a defect. To permit rapid correction of the defect, the prime equipment must be made up of replaceable plug-in assemblies. If this is not the case, it is academic whether the untrained technician can localize the defect to a circuit because when he finds it, he will not have the training to enable him to take corrective measures. However, if he localizes the defect to a replaceable assembly, he should be able to put the equipment back into operating condition quite readily. A major goal of this program was to develop a general purpose test set concept that would not require the technician to have any knowledge of the kind of test that he was making. The best way to accomplish this was to provide him with an instrument that has no controls to manipulate, does not require him to make any diagnostic decisions, and employs an identical test routine regardless of the variety of tests.

The intent of this development has never been to eliminate the need for the highly trained technician. Instead, the intent has been to improve the efficiency and effectiveness of maintenance procedures by permitting untrained technicians to assume relatively routine tasks and free the trained technicians for more difficult duties and/or to permit a reduction in the number of highly trained technicians required.

Throughout the development of the FIST techniques, strong emphasis was placed on the human engineering aspects of the maintenance problem. Frequently, otherwise adequate technical ideas were put aside because it was felt that the procedures required would evoke an unsatisfactory psychological response from the technician.

Many designers will find some of the techniques so attractive that they will be tempted to try retrofitting these techniques into existing equipment. In the initial stages of the program, a basic decision had to be made as to whether the techniques to be developed should be so constituted as to permit retrofitting. It was decided that the restraints resulting from a retrofitting requirement were so severe that novel and imaginative approaches to maintenance would be barred from consideration. Since this was a techniques program and the intent was to advance the state of the art, it was decided that no artificial liminations would be imposed. From the beginning of the program, the assumption was made that the techniques developed would be the type that must be designed into the prime equipment during the initial stage of the prime equipment development.

This then permits a certain latitude in the physical design of the prime equipment. The equipment obviously must be made up of replaceable assemblies. Normally the decision as to how an equipment should be divided into replaceable assemblies is made on the basis of assembly size. However, when applying FIST techniques to an electronic equipment, the circuits must be grouped into assemblies that can be efficiently tested by FIST techniques. For example, two tandem circuits in an equipment might be difficult to test individually but might be very efficiently tested together. Such circuits should be packaged into one assembly rather than two assemblies. Conversely, a large assembly that is difficult to test as a unit might be tested more easily if divided into smaller assemblies.

A prime equipment that has been designed for FIST maintenance must have adequate panel space on which to mount the multiple test sockets into which the general purpose test set can be plugged. These sockets should not be mounted on an exposed front panel but on a protected, easily accessible surface.

## THE BASIC BUILDING BLOCK

The basic keystone of the FIST maintenance system is the two-port measuring device, the comparator. A voltmeter is a one-port measuring device, while a comparator is a two-port measuring device. A two-port measuring device can be used to make one-port measurements but a one-port measuring device cannot be used to make two-port measurements. The two-port measurement is a versatile and powerful tool that unfortunately is not fully appreciated or exploited by electronic instrumentation engineers. Although few automatic checkout systems use comparators, because of certain restrictions placed on the systems, these comparators can rarely be used in an unconventional manner to locate faults.

## THE FIST COMPARATOR

The two-port measuring cell used in the FIST test system is a special phase- and waveform-insensitive voltage comparator (Figure I-1). For example, an AC signal applied to one port can be compared with a different kind of AC signal applied to the other port. A comparison made of peak-to-peak voltages is unaffected by variations of waveform or phase. For good sensitivity, each side of the comparator incorporates a wide-band amplifier.

The experimental FIST test set contains amplifiers with an upper cut-off of approximately 12 Mc. The outputs of the amplifiers drive peak-to-peak detectors. The DC outputs from these peak-to-peak detectors are compared in a DC differential amplifier. The null indicator in Figure I-1 is shown as a meter to permit the basic FIST principles to be most easily comprehended. Another useful null indicator is discussed in a later section which describes the test set details.

The restraints in many automatic checkout systems are so severe that they permit very few dynamic measurements to be made. This program emphasizes the other extreme and efforts were concentrated on dynamic measurement techniques since these are the most meaningful ones. Only occasionally can the function of an assembly be evaluated by measuring some key DC or low frequency AC voltage. More often the only way to assess the performance of a circuit is to measure its ability to provide the function for which it was designed.

Perhaps the simplest introduction to FIST principles is an examination of the technique for amplification measurement. The output signal of the amplifier in Figure I-2 is sampled by an attenuator comprised of two fixed resistors, $R_1$ and $R_3$, and one variable resistor, $R_2$. The attenuator output is connected to one input of the phase and waveform insensitive comparator and the input signal to the amplifier module being tested is connected to the other input of the comparator. The indicator on the comparator will null whenever the attenuation or loss is equal to the amplifier gain. In a practical case one desires to know whether the amplification is within the design limits, for example 9 to 11. For this example $R_1$, $R_2$, and $R_3$ would be so proportioned that an attenuation of 9 times would be realized when the arm of $R_2$ is at the upper end of $R_2$ and an attenuation of 11 times when the arm of $R_2$ is at the lower end of $R_2$.

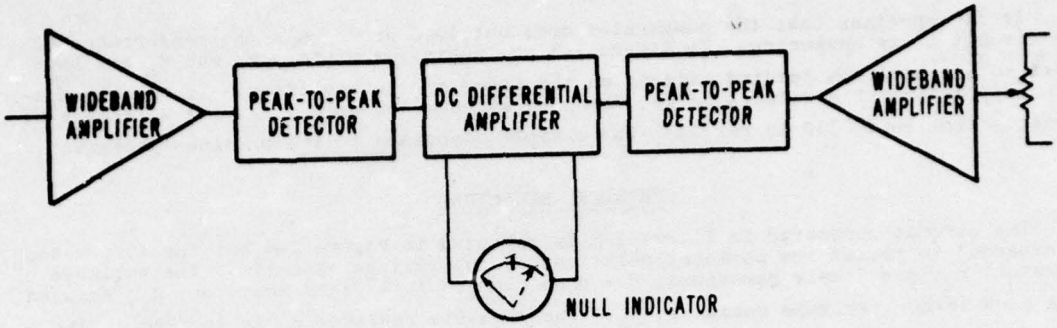
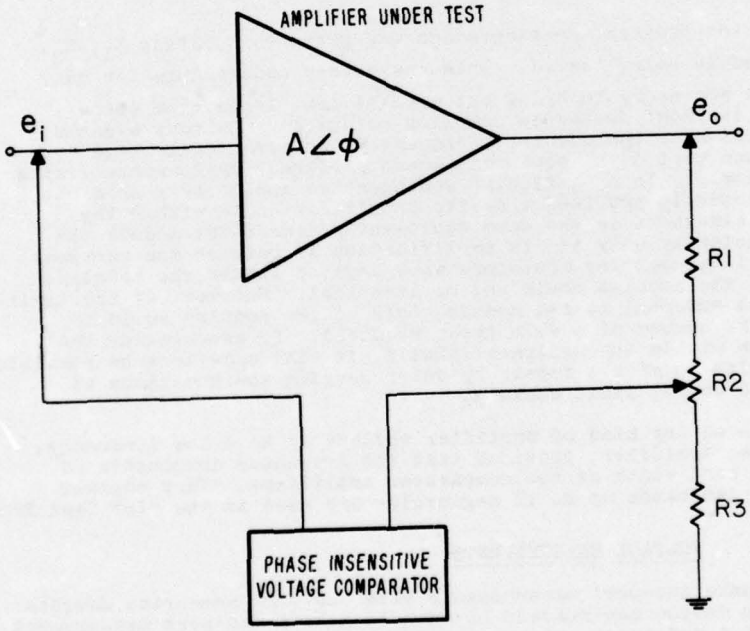



Fig. I-1 Basic Comparison Cell




$$\frac{R1}{R3} = A_{MAX}\left[1 - \frac{1}{A_{MIN}}\right]$$

$$R3 = \frac{R2}{\left[\frac{A_{MAX}}{A_{MIN}} - 1\right]}$$

Fig. I-2 Comparison Method

Thus, some position of the arm of $R_2$ will permit a null to be obtained if the amplification is within the design limits and conversely the inability to obtain a null for any position of the $R_2$ arm is an indication that the amplifier performance is outside of its design limits and has therefore failed.

It is important that the comparator does not load or degrade the performance of the circuit it is measuring. In Figure I-3 the resistive divider $R_4$ and $R_5$ has been added to minimize any loading effects on the input of the amplifier. If the attenuation supplied by this network is 10, then the attenuation of $R_1$, $R_2$, and $R_3$ should be variable from 90 to 110 to parallel the example discussed in the previous paragraph.

## PHYSICAL STRUCTURE

The circuit presented in Figure I-3 is repeated in Figure I-4 but the information is expanded to reveal the physical relations of the various elements. The variable resistor, $R_2$, previously described, has been changed to a fixed resistor, $R_2$, shunted by a much larger variable resistor, $R_2'$. The variable resistor $R_2'$ is located in the test instrument so that the operator can conveniently manipulate it. The test socket is located on a readily accessible panel surface. Since the socket into which a module plugs can rarely be located in a position adjacent to the test socket, short shielded cables will usually be required to convey the test signals to the test socket. $R_1$ and $R_4$ should be mounted on the module socket so that the cable capacitances will not load the module. $R_1$ and $R_4$ must be sufficiently large so that even if their lower ends become grounded, the normal function of the module will not be seriously affected. $R_2$, $R_3$, and $R_5$ should be located in a compact cordwood assembly mounted on the back of the test socket.

In the early phases of the program, consideration was given to locating $R_1$, $R_2$, $R_3$, $R_4$, and $R_5$ within the module being tested. This was deemed undesirable for many reasons. It would have been necessary to bring out special test leads from each assembly. It was felt that the most desirable sampling points for the test signals were at the input and output pins of the module. Frequently the same module type could be used in more than one part of an equipment where different performance limits would be tolerable. For example: In a particular equipment an amplifier module can perform its function satisfactorily provided that its amplification is within the limits of 5 and 11, whereas elsewhere in the same equipment an identical module can perform its function satisfactorily only if its amplification is between the narrower limits of 9 and 11. If the limit-setting resistors were located within the modules, then, for the example cited, the modules could not be identical. However, if the limit-setting resistors are located external to the module, both of the modules could be identical, thus minimizing the number of module types required. By associating the test limits with the position within the equipment itself, it will sometimes be possible for the knowledgeable technician to make a repair by interchanging the positions of identical modules in wide and narrow limit sockets.

Measurements can be made on any kind of amplifier whether it be a low frequency, high frequency, or pulse video amplifier, provided that the frequency components of the signal do not exceed the band width of the comparator amplifiers. Very compact gain stable amplifiers with pass bands up to 12 megacycles are used in the FIST Test Set.

## VOLTAGE MEASUREMENTS

It is not practical to make two-port measurements with one-port measuring devices; however, a two-port measuring device can readily be used to make a one-port measurement by comparing the unknown with a reference voltage (Figure I-5). Figure I-6 illustrates a practical transformation network technique that would be used to make a voltage measurement. The voltage to be measured may be any periodic wave form that does not have substantial frequency components outside of the comparator pass band. The reference voltage, contained within the test instrument, may be any AC signal of known amplitude. A square wave reference voltage may be conveniently obtained by regulating a DC voltage with a Zener diode and then using this voltage to power a transistor multivibrator whose

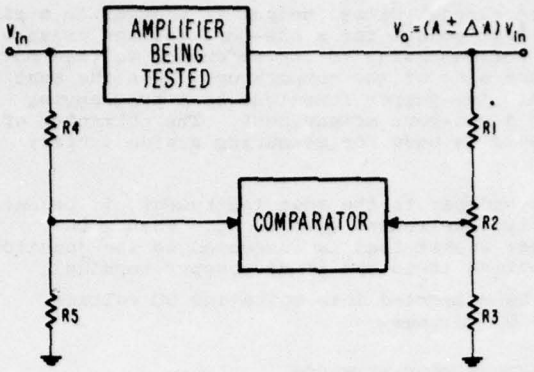



Fig. I-3 (above) Practical Circuit For the Measurement of Amplification

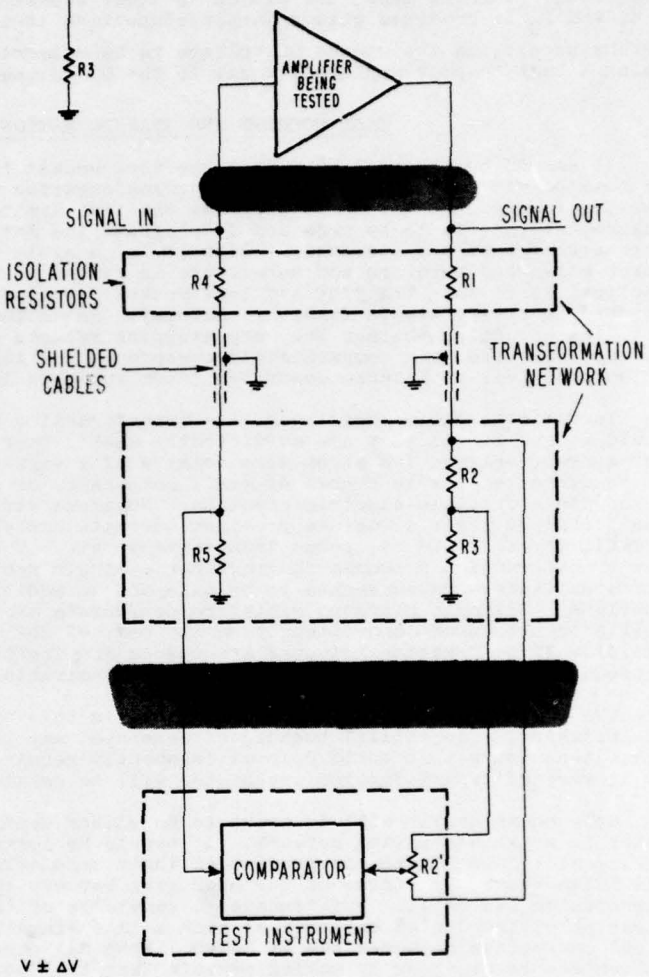
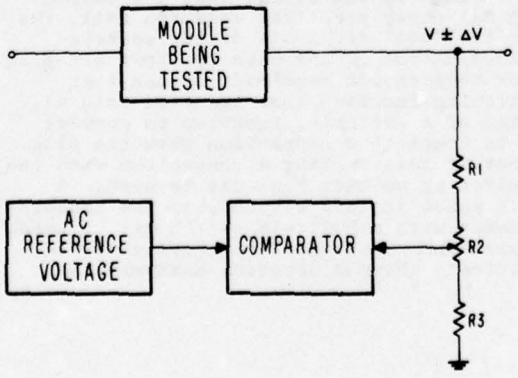



Fig. I-4 (right) Physical Relationship of the Components in a One-Cell Test

Fig. I-5 (below) Basic Voltage Measurement



transistors bottom when turned on. The AC reference voltage output is brought to a pin in the test instrument plug. To set up the test socket for a one-port voltage measurement, a jumper is connected between the pin corresponding to the reference voltage output and the pin corresponding to the reference side of the comparator. When the test instrument is plugged into such a test socket, the jumper functions as a programming connection to set the test instrument up for a one-port measurement. The character of the comparator permits one reference voltage to be used for measuring a wide variety of wave forms.

Figure I-7 shows how, by adding a shunt chopper to the test instrument, it becomes possible to extend the measurement range of the instrument down to DC. When a DC measurement is to be made, the pin on the test socket that is connected to the junction of $R_1$ and $R_2$ is provided with a jumper connecting it to the shunt chopper terminal, thereby permitting the steady DC voltage to be converted into pulsating DC voltage having a peak-to-peak amplitude equal to the DC voltage.

## TEST SOCKETS AND TRANSFORMATION NETWORKS

It should be observed here that the test socket is not just a convenient terminus for test points but, together with the transformation network, supplies a test programming function as well. It programs the test limits, it programs the kind of measurement that is to be made and it programs the interconnection of the general purpose FIST test set elements. While it is possible to eliminate the plug and test socket selection hardware and substitute an automatic selecting mechanism, it is not practical to do so. The plug and test socket permit a great many connections to be switched in a very simple fashion. Automatic selection devices having the same switching capacity, whether they are stepping relays, crossbar switches or semiconductor devices, are comparatively cumbersome and impractical. In addition, most of the practical advantages described later would be lost.

To maximize their usefulness, the transformation network assemblies (Figure I-4) should be small. If they are sufficiently small, they can be mounted directly on the test socket. Figure I-8 shows some details of a versatile cordwood structure that can accommodate a large number of small components or a few large components such as RF inductors or piezo-electric crystals. Numerous versions of these assemblies are possible since their structure provides adequate space to accommodate components such as small crystal holders, pulse transformers, etc. The assemblies in the illustration show transformation networks ranging from a single resistor to a network that permits four simultaneous measurements to be made and in addition can perform switching functions. Although standard miniature components are used, the assemblies are still small enough to mount conveniently on the rear of the test socket. Until such time as thin film transformation networks are common and inexpensive, these standard component cordwood assemblies should prove to be quite desirable because of their small size.

The transformation techniques described in this and later sections are far from all-inclusive. The initial techniques developed may be considered to be the "work horse" techniques that would be most frequently required. In time it is expected that the library of transformation techniques will be considerably expanded.

Only occasionally will it prove to be either desirable or necessary to include a switch in a transformation network. It should be borne in mind that many of the test sockets will have AC signals present at their terminals and occasionally a situation will arise where, by virtue of the high gain between two sockets, cross talk and/or regeneration can exist. Unfortunately, knowledge of this difficulty comes to the attention of the design engineer at such a late stage in the design that few conventional corrective measures can be taken. When all other practical measures fail, the problem can be overcome by making certain that the "hot" terminals of the sockets involved are normally grounded and are disconnected from ground when the test set plug is inserted into the socket. Although a jumper between two terminals of the test instrument plug is capable of performing a switching function when inserted into a socket, unfortunately it performs the wrong kind of a switching function to correct the cross talk problem. The jumper functions to complete a connection when the plug is inserted, whereas the required action is that of interrupting a connection when the plug is inserted. There are three possible switching methods that can be used. A jumper on the test set plug can actuate a relay which in turn disconnects the troublesome socket terminal from ground; a special socket with a built-in switch may be used, or a conventional socket may be modified to accommodate a standard microswitch. The latter solution was deemed to be the most practical. Simple mounting hardware was

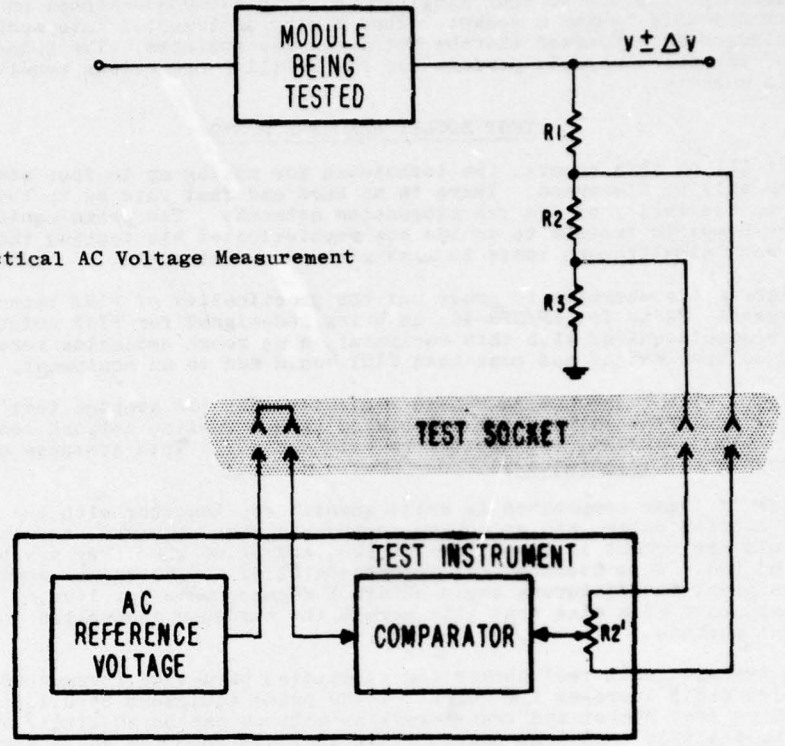



Fig. I-6 Practical AC Voltage Measurement

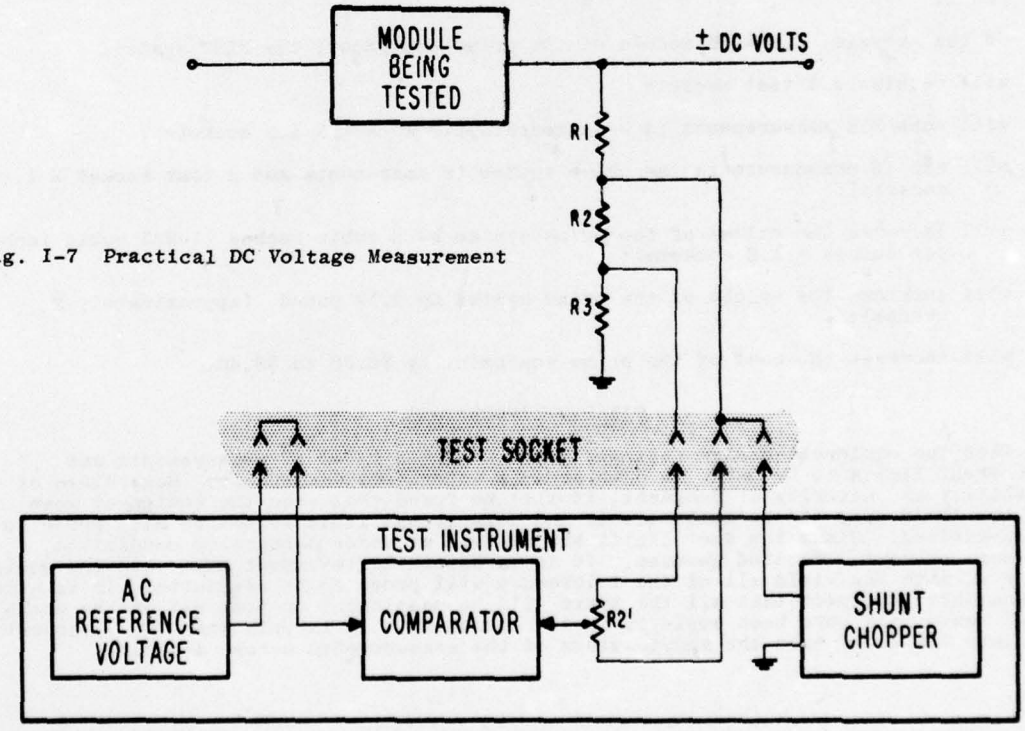



Fig. I-7 Practical DC Voltage Measurement

developed for mounting one to four single-pole, double-throw switches on the transformation network assembly behind a socket. When a plug is inserted into such a modified socket, a plunger is depressed thereby actuating the switches. The prime equipment manufacturer can inexpensively perform the few drilling operations required to modify the standard sockets.

## TEST SOCKET HARDWARE COSTS

In Part III of this report, the techniques for making up to four simultaneous measurements will be discussed. There is no hard and fast rule as to the degree of simplicity or complexity of the transformation networks. The prime equipment design engineer has complete freedom to decide how sophisticated his testing techniques should be and how many simultaneous tests to make with each test socket.

As a laboratory exercise to prove out the practicality of FIST techniques, a simple equipment, Radar Set AN/SPS-46, is being redesigned for FIST maintenance. Based on the experience acquired with this equipment, some rough estimates were made of the complexity, volume, weight and cost that FIST would add to an equipment.

Experience with Radar Set AN/SPS-46 indicates that the average test socket is used to make three measurements, and the associated transformation network contains nine components - six resistors, two capacitors, and a diode. This averages out to only three components per measurement.

The cost of these components in small quantities, together with the test socket, one foot of coaxial cable, and switching components when required, totals $7.00 when the components are bought in quantities of 100, and $5.00 when they are bought in quantities of 500. This averages out to between $1.67 and $2.33 per measurement. Original equipment manufacturers could obtain the components for less. This cost is based on components of a size that will permit the cordwood assemblies to be removed with the test sockets.

On the average, each test socket and associated hard-ware (transformation network, coaxial cable, etc.) increase the weight of the prime equipment by 0.1 pound. The volume occupied by the test socket and transformation network can be as little as 1-1/3 cubic inches behind the 1/16 inch test panel. This includes room for cable runs behind the transformation network. The test socket also requires 1/3 cubic inch in front of the test panel.

On the average, for each module of the prime equipment, the FIST system:

will require 1.2 test sockets;

will make 3.6 measurements (3 measurements per socket X 1.2 sockets);

will add 12 components to the prime system (9 components and a test socket X 1.2 sockets);

will increase the volume of the prime system by 2 cubic inches (1-2/3 cubic inches per socket X 1.2 sockets);

will increase the weight of the prime system by 0.12 pound (approximately 2 ounces);

will increase the cost of the prime equipment by $6.00 to $8.40.

## FIELD MODIFICATIONS

When the equipment design engineer decides on the types of measurements and measurement limits to be used, he does so in a laboratory atmosphere. Regardless of his ability and maturity of judgment, it will be found that when the equipment goes into the field some of the measurements and measurement limit decisions will prove to be ill-advised. Since the test limits are arrived at under laboratory conditions, they are, at best, educated guesses. It is unreasonable to expect that when the equipment goes into the field all of the tolerances will prove to be realistic. It is also unreasonable to expect that all the tests will be meaningful or that all of the meaningful tests will have been employed in the equipment. It is only when the equipment gets into the field that the shortcomings of the measurements become apparent.

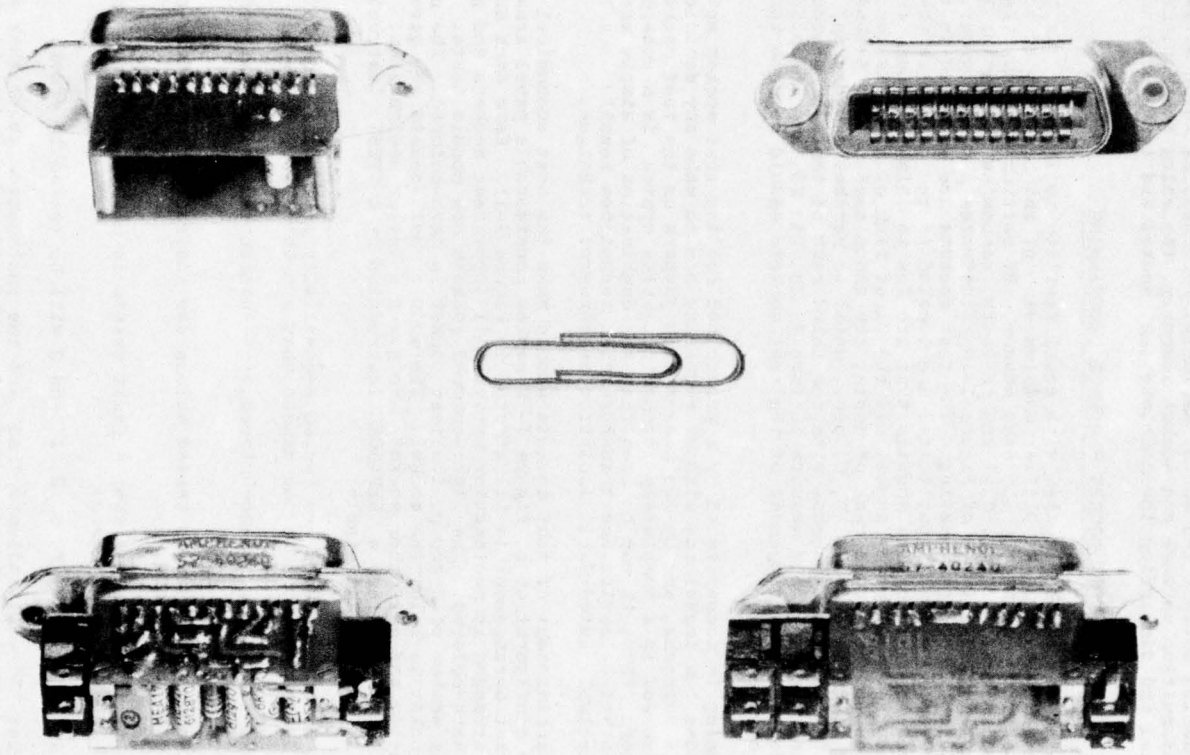


Fig. I-8 Test Sockets and Transformation Networks

In most automatic testing systems it is difficult and frequently impossible to make a field modification if hardware must be replaced. Figure I-9 shows a typical FIST field modification kit, consisting of a replacement socket and transformation network. The technician can readily update a maintenance test in the field by removing the screws that fasten the test socket to the panel, unsoldering a few wires from the obsolete transformation network and socket assembly; reconnecting these wires to the replacement transformation network and socket assembly, fastening the replacement socket to the panel and discarding the obsolete test socket and transformation network.

## TEST SOCKET SEQUENCE PROGRAMMING

The FIST techniques can be applied with equal facility to large or small aggregates of circuits. For example, three amplifier modules each of which is tested individually, can be tested as a group in exactly the same manner. By setting up group tests in a piece of equipment as well as individual module tests, economies in testing time can be realized. The test socket display of Figure I-10 illustrates a socket grouping order that permits efficient and rapid testing. The test sockets are arranged in three ranks. The first rank of tests, signal input tests, are located in the column on the left. These determine the adequacy of the signals that are fed to this equipment from other sources as well as the prime power input. If the first rank of tests is good, the technician proceeds to the second rank of tests, the group tests. The technician goes through the group tests in order A, B, C, etc., until he reaches a bad group test (for example E). At this point he proceeds with the third rank of tests, the module tests, and in the example cited he would measure in turn 4, 23, 24, 25, and 15 until he locates the defective module. The arrangement of the test sockets should be such that the testing order is obvious.

FIST programming is accomplished by a prescribed routine test socket selection sequence that guides the technician without requiring him to make any decisions, by the transformation networks, and by the programming jumpers on the test sockets. To accomplish the same end in a completely automatic testing system is a considerably more complex matter. Thus, it may be seen that the combination of simple manual operations involving socket selection and transformation techniques results in a basically simple and readily implemented fault isolation measurement technique.

Ideally the arrangement of test sockets should make the most economical use of panel space. The configuration of Figure I-10 wastes considerable panel area. A superior test socket arrangement is illustrated in Figure I-11. Here dark and light test sockets are arranged in rectangular array. All input test sockets and group test sockets are dark-colored. The light-colored sockets are module tests. If a module is the only member of a group, its test socket is dark-colored. The nomenclature of each socket is below the socket. The module test sockets are given a numerical designation and all other sockets are given a letter designation. Imprinted above each socket will be either a "REPLACE" instruction or a "TEST" instruction. The testing sequence is uniform and simple:

1. The dark-colored sockets are tested sequentially proceeding from left to right and top to bottom in the manner that a printed page is read.

2. When a bad indication is encountered, the instruction above that socket is followed.

3. No light-colored socket is tested unless the instruction on a dark-colored "bad" socket so directs.

For example, if it is assumed that a fault exists in module A-26, the following test routine will automatically ensue:

1. Sockets A, B, C, D, E, F, G, H, I, and J will be tested in order.

2. Test socket "J" will indicate "bad" and the instruction above that socket will direct the technician to light-colored sockets 15, 16, 17, 18, 19, 20, and 21.

3. By following these instructions the technician will find that test socket "20" indicates "bad." The instruction, located above that socket, will direct the technician to replace module A-26.


Fig. I-9 Field Modification Kit

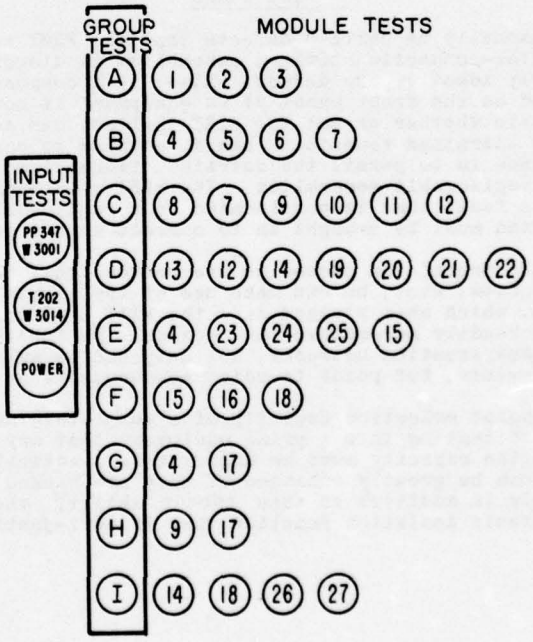



Fig. I-10 Test Sequencing by Socket Positioning

It is never necessary to read the socket nomenclature unless a bad indication is observed. Occasionally more than one test socket is required to completely test a module. This need be no concern of the technician, since his instructions for corrective action are automatically conveyed to him when he observed a failure on a test socket.

The testing sequence is set up so that the first tests are always input tests (Figure I-11, sockets A, B, C, and D). The instruction above the group test sockets directs the technician to the test sockets of the modules that comprise the group. When there is only one module in a group its test sockets will be dark-colored.

It is not necessary for the technician to be acquainted with the design rules that lie behind the test socket arrangement since he is required only to follow an invariant testing procedure.

## SUGGESTED PHYSICAL STRUCTURES FOR FIST-MAINTAINED EQUIPMENT

The panel surface on which the test sockets are mounted should be chosen to be readily accessible during a testing sequence yet provide protection for the sockets between maintenance periods. Figure I-12 is suggested as one way these objectives can be met in modularized, conventional, rack-mounted equipment. If the equipment is mounted on slides that may be tilted 90°, the entire lower surface of the chassis becomes a convenient plane on which to locate the test socket panel. When mounted in this manner, there is no tendency for the test sockets to collect debris or moisture between maintenance periods.

In the case of micro-modularized equipment it is expected that each replaceable assembly will contain more than the usual amount of circuitry, so that the size of these replaceable assemblies will be sufficiently large to warrant the allocation of space for FIST test sockets and transformation networks. A suggested physical structure for microminiaturized equipment is illustrated in Figure I-13.

The structures in Figure I-12 and I-13 were both devised with the intent of keeping the shielded cable runs between the test sockets and their modules as short as possible. Unless only low frequency and DC voltages are being measured, it would not be practical to locate the test socket panel at a point remote from the modules being tested and use long shielded cable runs because of their capacity. This would rule out certain aircraft applications where one might be tempted to place the test socket panel on the external skin of the aircraft.

## FIST USAGE

There will occasionally be certain defects that the FIST test set cannot identify. For example, if an inter-connection between assemblies is discontinuous, the FIST test set will not positively identify the defect. Also, if a component such as a potentiometer which is mounted on the front panel of an equipment is not part of a replaceable assembly, it is academic whether or not the FIST test set can identify that it is defective because the untrained technician can do nothing to correct the problem. The goal of FIST maintenance is to permit the untrained technician to isolate only those faults that exist in replaceable assemblies. The FIST maintenance procedure is not designed to pinpoint a fault that is not located in a replaceable assembly. A knowledgeable technician must be brought in to correct such a defect.

Should the trained technician choose to use conventional test equipment such as multimeters, oscilloscopes, etc., he can make use of the simple test adapters, illustrated in Figure I-14, which when plugged into the FIST test sockets will provide him with a vast number of readily accessible test points. By reading through the resistive attenuators in the transformation networks, not only can he make voltage, waveform and signal tracing measurements, but point-to-point measurements as well.

The finite test point selection capacity of a completely automatic test system determines the depth of testing into a prime equipment that may be accomplished. Since this test point selection capacity must be kept within practical bounds, the usefulness of automatic systems can be greatly enhanced if they are backed up by the FIST maintenance system. Obviously in addition to this backup ability, the FIST maintenance system is able to furnish a fault isolation function that is self-justifying.

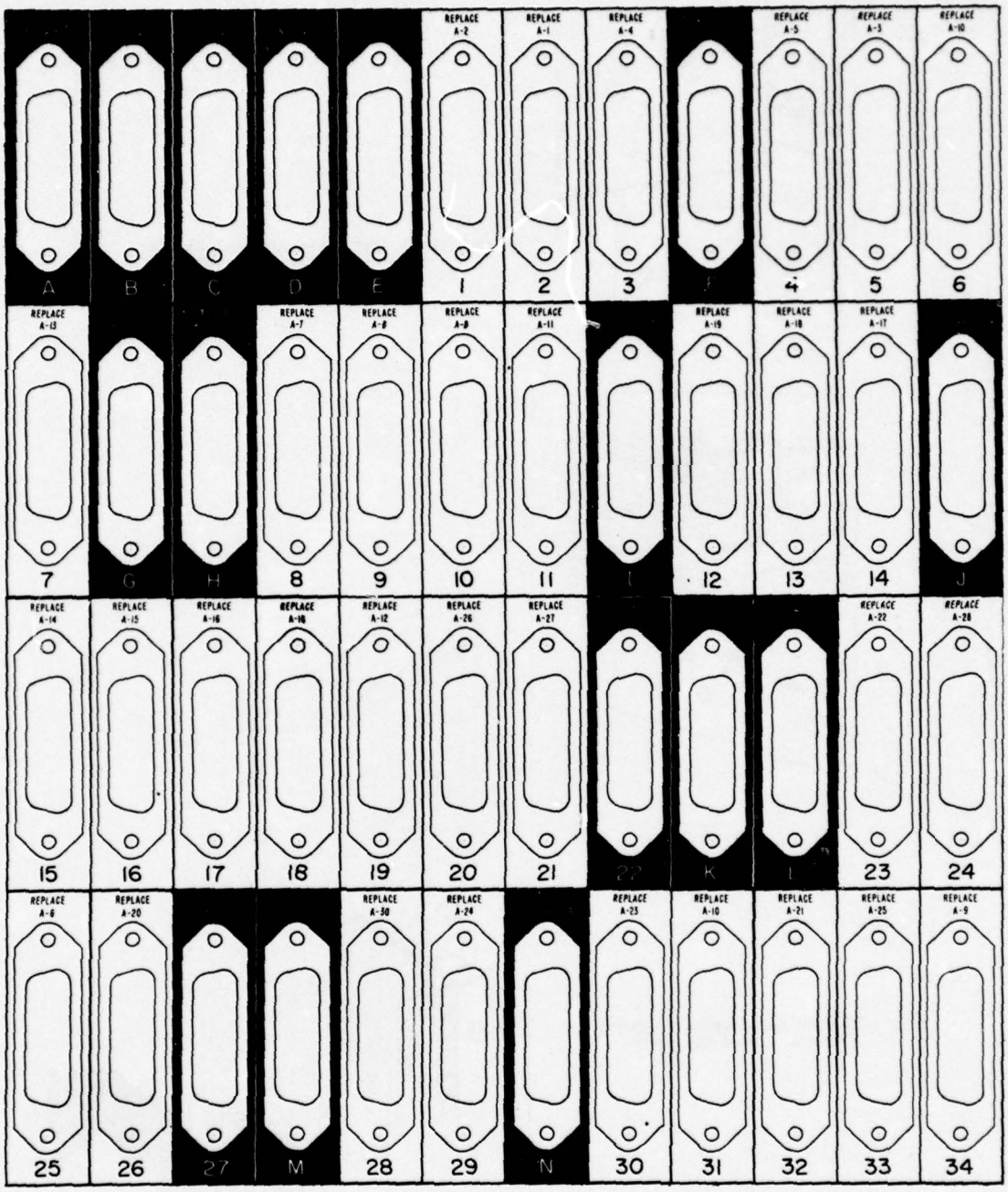



Fig. I-11 Recommended Test Socket Arrangement and Instruction Markings

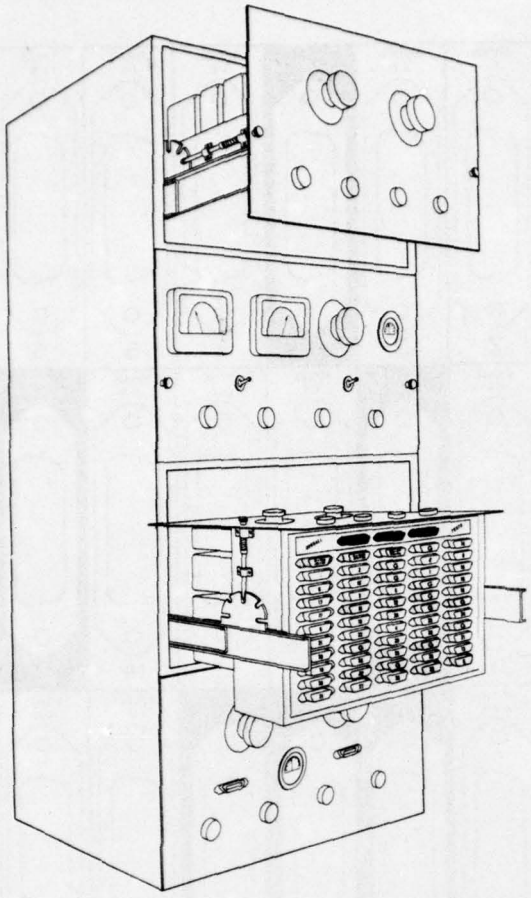


Fig. I-12 FIST Application to Rack and Panel Mounted Equipment

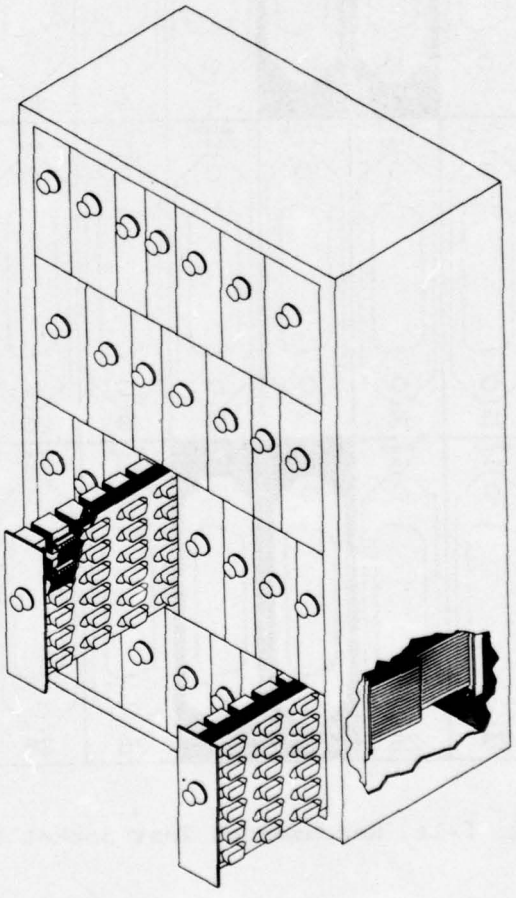
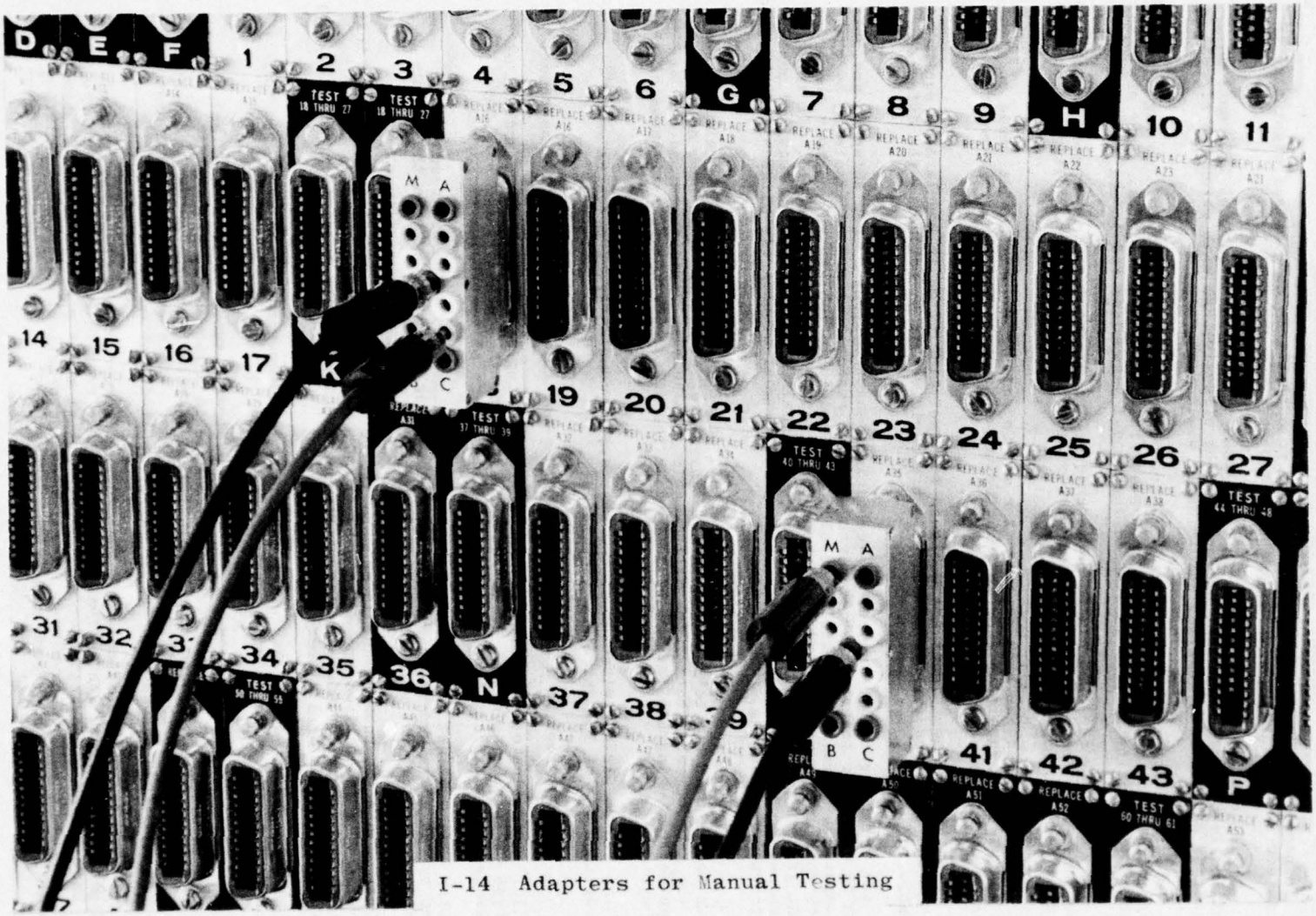


Fig. I-13 FIST Application to Micro-miniaturized Equipment




I-14 Adapters for Manual Testing

## A SUMMARY OF FIST OBJECTIVES AND FEATURES

To permit a better understanding of the role that this development might play in the future, it is appropriate to discuss its objectives and features:

(1) Emphasis has been placed on tests that will yield meaningful information, specifically dynamic tests that can measure the effectiveness with which the circuit under test performs the function for which it was designed.

(2) These dynamic tests are general in nature, equally applicable to high-power transmitting vacuum tube circuits or low-power transistorized circuits. They are equally capable of isolating defects in simple, functional assemblies and in complex assemblies.

(3) Emphasis has been placed on techniques for diagnosing non-computer faults, since there exist programming and test problem techniques that are peculiarly useful for diagnosing computer malfunctions.

(4) The system has been designed for set-up and testing time under field conditions that is orders of magnitude shorter than the set-up and testing time required for conventional test equipment.

(5) The FIST techniques are designed for fault isolation in modularized equipment so that the untrained technician can effect a rapid repair by replacing the defective assembly.

(6) Faults which the untrained technician cannot correct by replacing an assembly (such as defective interconnection wiring between assemblies) are not isolated by the test system. A trained technician should be called in to isolate and correct those faults that cannot be remedied by the simple action of replacing a module. It is the intent of this program to achieve the practical goal of minimizing the required number of trained technicians, not to seek the impossible goal of entirely eliminating the need for trained technicians.

(7) Although the system has been devised chiefly for use by the untrained technician, who should not be required to make any decisions, it has also been designed so that the knowledgeable technician will be able to make very rapid tests and not be prevented from using his initiative to improve the efficiency and effectiveness of the test system.

(8) To permit minimization of the test system complexity, the system seeks to make optimum use of the untrained technician.

(9) The test set is a general-purpose, man-portable, semi-automatic instrument.

(10) Operation of the test instrument is very simple and uniform, regardless of the type of test being made. It is not necessary for the technician to have any knowledge of the circuit type being tested or the type of test being performed.

(11) The test set incorporates self-testing techniques which are readily comprehended by the untrained technician, thereby enhancing his confidence in the proper functioning of the test set when the self-test so indicates.

(12) Since the human being is so constituted that he most effectively employs that to which he is favorably disposed, the system has been human-engineered to impart to the technician a ready understanding of its basic principles, a trust in its reliability, and an appreciation of its operation ease.

(13) Simple but efficient test-programming techniques have been developed to obviate the necessity for testing every module.

(14) Engineering decisions regarding the desirable types of tests and test tolerances required by a prime equipment are made in a laboratory atmosphere. Occasionally it will be found that cumulative field reports on the prime equipment will dispute the wisdom of the original testing choices. It is practical to take remedial action in the field through the use of easily applied test modification kits.

(15) The techniques advanced by the program are not of the "blue-sky" variety, but depend upon current state-of-the-art technology that is ready to be used here and now. Many of the engineering hardware problems that might prevent successful application of the techniques have been investigated and practical engineering solutions to these design problems have been developed.

(16) The application of the testing techniques is relatively simple and will be codified in the form of a design manual so that the prime equipment design engineer can incorporate the fault location features in his equipment without requiring either the intervention or cooperation of the test equipment specialist. The intent here is to eliminate the interface problem that exists between the prime equipment designer and the automatic test equipment designer.

# PART II

## TRANSFORMATION NETWORKS

Table II-1 Properties that can be Measured by the FIST System

- AMPLIFICATION
- BANDWIDTH
- CAPACITANCE
- CURRENT
- FREQUENCY
- IMPEDANCE
- INDUCTANCE
- PHASE SHIFT
- PHYSICAL*
- POWER
- PULSE
  - AMPLITUDE
  - DECAY TIME
  - DELAY
  - DROOP
- PULSE (continued)
  - DUTY FACTOR
  - OVERSHOOT
  - RATE OF RISE OR DECAY
  - REPETITION RATE
  - RISE TIME
  - WIDTH
- RADIATION (heat, light, x-ray, etc.)
- RESISTANCE
- SAWTOOTH LINEARITY
- SQUARE WAVE SYMMETRY
- TEMPERATURE
- VOLTAGE
  - AC (with or without dc component)
  - DC

* A variety of physical properties (mechanical, hydraulic, etc.) can be measured using physical-to-electrical transducers.

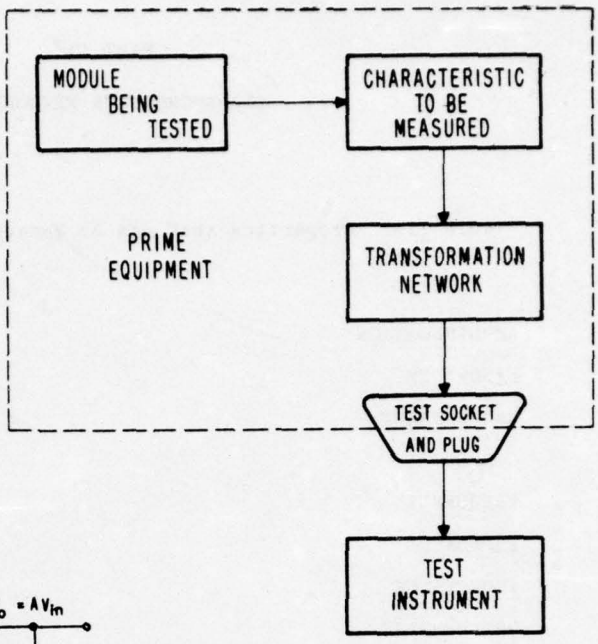



Fig. II-1 (right) Functional Diagram Showing the Relationship Between the Module Being Tested and the Parts of the FIST Measurement System

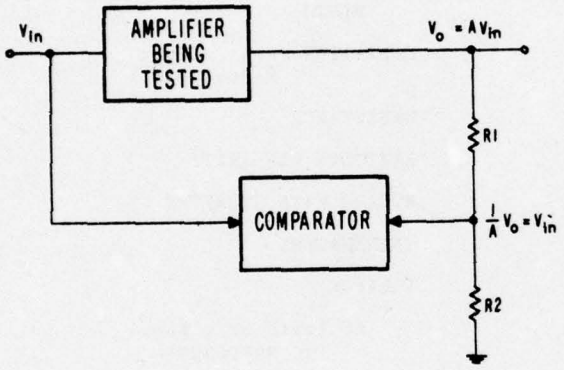



Fig. II-2 (left) Basic Circuit for Measuring Amplification

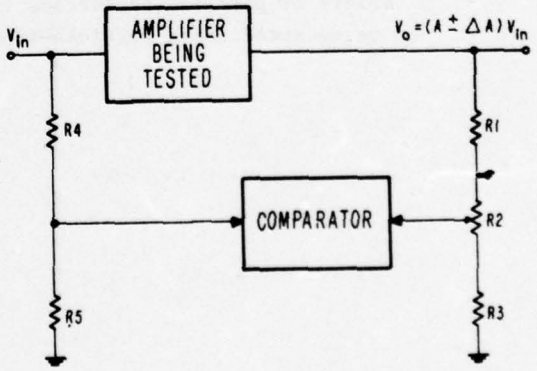



Fig. II-3 (right) Practical Circuit for the Measurement of Amplification

# PART II

# TRANSFORMATION NETWORKS

## INTRODUCTION

Transformation networks are the electrical networks used to convert a critical parameter of the prime system under test to a signal that is acceptable by the FIST test instrument. Physically, the transformation network is a part of the prime system and forms the interface between the prime system and the test instrument. It includes all the components added between the point on the prime equipment at which the system parameter is sampled and the test socket. Most of these components, which are usually passive elements, are mounted in a cordwood assembly behind the test socket.

The FIST system is used to determine whether the performance of each module of an electronic system is within the limits required for the correct functioning of the system. This is usually done by measuring the dynamic performance characteristics of the module, but it may also require measuring DC currents and voltages, critical properties of the environment within the module, and occasionally values of resistance, inductance, or capacitance. The transformation networks provide the versatility that permits the FIST system to measure these characteristics using a test instrument that is basically a device for comparing the peak-to-peak amplitudes of two periodic voltage waveforms.

For simplicity of operation and speed, the FIST system is designed so that the measurements can be made without reference to technical manuals, schematic or block diagrams, trouble shooting charts, resistance or voltage tables, waveform drawings, etc., even though the operator is not familiar with the equipment being tested. The information that would normally be obtained from these sources is designed into the transformation network. This information includes the selection of the module characteristics to be measured, and the setting of the acceptance limits of the test based on the permissible limits of the characteristic being measured.

The transformation network converts the characteristic being measured to a voltage whose amplitude is a sensitive index of small percentage changes in the characteristic. A voltage divider included in the transformation network reduces this voltage to the level required by the test instrument, and sets the acceptance limits of the test based on the requirements of the module being tested. The functional relationship between the transformation network, the module being tested, and the test instrument is illustrated in Figure II-1.

Making the transformation network an independent part of the system has several advantages:

First, placing the transformation network outside the module permits testing only those properties of a standard module which are critical in the particular application. Not all of the characteristics of a standard module are critical in every application. For example, if the output of the module is a pulse, the rise time and amplitude may be the most important properties if the output is used for triggering, whereas the width and amplitude may be the critical characteristics if the output is used for display. Since, in the case of a standard module, the application will not be known in advance, it would be necessary to provide for testing all the output characteristics of the module if the transformation network were placed inside the module. Placing the transformation network outside the module permits testing only the essential characteristics and therefore usually results in a simpler transformation network.

Second, placing the transformation network outside the module permits testing the characteristics of the module to the specifications required by the application. A module can be used in any application in which the performance requirements are not more stringent than the acceptance requirements for the module. The only practicable way of testing the modules in such applications is to base the test on the actual and frequently broader limits required by the application. This requires placing the transformation network outside the module, since the limits will be different for each application.

Finally, since testing the module to the limits required by the application results in a large number of different transformation networks, the only practical solution is to make the transformation networks part of the prime equipment, rather than of the test instrument. While it is true that the transformation network could be simplified by providing ranges on the test instrument rather than by normalizing the voltage within the transformation network, the former would require the use of supplemental charts or manuals which would slow up the testing.

A prerequisite for design of a transformation network is a knowledge of the characteristics of both the system under test and of the test equipment being used. Once the characteristics of the FIST test set are standardized, however, the transformation network configuration required for each type of measurement can be established so that only the actual component values must be determined to test the magnitude and limits of the characteristic to be measured in a specific application. The use of nomographs and similar aids should simplify the problem of designing the transformation networks into the prime equipment.

A handbook giving the details of transformation network design for determining component values is planned. These details are not included in the present report, which is intended to show the variety of measurements that can be made using a comparator-type test instrument (see Table II-1) and to illustrate a circuit configuration which can be used to make each type of measurement. It should be remembered that there is usually more than one way of converting the characteristic to be measured to a voltage that is within the range of the test instrument. No attempt has been made here to illustrate all the possible ways in which this can be done. The transformation networks described in this section are among the simplest that can be used to provide the desired information with an acceptable degree of accuracy.

The test instrument for which these transformation networks are designed is basically a device for comparing the peak-to-peak amplitude of two periodic voltage waveforms. Since the test set operates on peak-to-peak amplitudes, the two inputs need not be in phase nor have the same waveform. DC signals are measured by converting them to AC by means of choppers. The test instrument sometimes receives one and sometimes both of its input voltages from the module under test. When only one input voltage is received from the module, the second is supplied by an internal reference voltage within the test instrument. By means of a meter reading or indicator lamps, the test set registers "good" when the amplitudes of its inputs are equal, "bad" when they are not, and "no-test" when the stimuli to the module are outside their prescribed limits. For speed and simplicity of operation, there are no operating controls on the test instrument.

## AMPLIFICATION

1. General:

Amplification is easily measured dynamically. The method of measurement is independent of the type of signal being amplified so long as it is DC or periodic in nature and within the frequency range of the test instrument. For this measurement, both inputs to the comparator are derived from the circuit being tested. The signal present at the input to the amplifier is used as the basis for the comparison signal, while the output signal furnishes the test signal. The transformation network in this case is a simple voltage divider or a pair of voltage dividers.

The test for amplification is shown in its simplest form in Figure 1I-2. The input signal to the circuit under test is also used for the comparison input to the comparator, while a fraction of the output voltage of the amplifier being tested is used as the test input. If the nominal amplification of the circuit is A, and the fraction of the output voltage which is fed to the comparator is 1/A, the test instrument will indicate a balance only when the amplification of the circuit under test is the specified nominal value. In most cases this test can be made while the amplifier is performing its normal function in the system. As long as the amplifier is operating within its linear region, the test is independent of signal level.

A more practical circuit is illustrated in Figure II-3. This transformation network permits a good indication to be obtained when the amplifier under test is operating within any specified limits of amplification. Voltage divider, $R_4$,$R_5$, is used to minimize the loading on the input of the circuit under test, and also, if necessary, to

permit reducing the signal voltage present at the module input to the level required by the comparator. Divider, $R_1, R_2, R_3$, reduces the output voltage of the amplifier under test to the level of the signal at the comparison input for application to the test input of the comparator.

If the input divider, $R_4, R_5$, reduces the module input signal by a factor n before it is applied to the comparator, the output divider, $R_1, R_2, R_3$, must reduce the output voltage by a factor nA where A is again the amplification of the circuit. The resistance of potentiometer, $R_2$, relative to the total resistance of the divider, is chosen so that the range of output voltages which can be obtained from the divider will permit the inputs to the comparator to be balanced only when the amplification of the circuit under test is within the acceptable limits of its performance.

If the amplification of the circuit under test has dropped to its minimum acceptable value, a larger fraction of the amplifier output voltage will be needed to balance the comparison signal from the amplifier input. If the resistor values are chosen so that the potentiometer arm is in its extreme upward position when this amplification occurs, a balance can be obtained for the lowest acceptable limit of amplification but not if the amplification drops any lower. Similarly, if the amplification is at its highest acceptable value, the output voltage of the divider must be a smaller fraction of the total, and balance should be obtained at the most downward position of the potentiometer arm. Higher amplification will result in an output voltage which is too high to permit a balance within the range of the potentiometer, and the module will again test bad. In other words, the amplifier is acceptable only if a balance can be obtained within the range of the potentiometer.

For example, assume that the performance of an amplifier is acceptable provided its amplification is between 9 and 11. The input signal to this amplifier is 10 times the maximum signal level required by the test instrument so that input divider, $R_4, R_5$, must reduce this signal by a factor of 10 to provide the comparison input to the comparator. To match this signal at the test input, the output divider, $R_1, R_2, R_3$, must attenuate the amplifier output signal by a factor of 90 when the amplification is minimum (test input to the comparator is then $1/90 \times 9V_{in} = 0.1V_{in}$) and by a factor of 110 when the amplification is maximum. If the output divider is designed so that its output is 1/90 of its input when the potentiometer arm is in its uppermost position, and 1/110 of its input when the potentiometer arm is in its most downward position, the potentiometer can deliver any fraction of the amplifier output voltage between 1/90 and 1/110 to the test input of the comparator. This permits balancing the inputs to the comparator for any amplification between 9 and 11. The limits to which the amplification is checked can be changed by changing the potentiometer limits.

Since the "good - bad" indication is independent of the relative phase of the two input signals to the comparator, the test for amplification is not affected by the phase shift of the amplifier under test, and is the same whether this amplifier inverts the signal or not. If the amplification of the circuit is less than unity, such as occurs in a cathode follower, the amplifier input must be attenuated more than the output, but in other respects the test is the same. In all cases the total resistance of the input and output dividers must be high enough to avoid loading the circuits to which they are connected, but low enough to prevent shunt capacitive reactance from introducing significant error in the voltage divider.

It is apparent that an erroneous indication could result if there were no input to the amplifier at all. In this case the comparator would indicate a balance because the two inputs would be equal, in this case zero, and the amplifier would apparently test good when, in fact, it was not being tested. This ambiguity is removed by simultaneously measuring other properties of the module or of its stimuli and using these as "qualifying measurements." The test instrument is programmed so that these qualifying measurements permit a "good" or "bad" indication to be obtained if they are all correct, but cause a "no test" indication when they are not. In the present instance, the amplitude of the input signal to the amplifier might be used as a qualifying measurement. This qualifying test is programmed in such a way that if the input signal amplitude is within the prescribed limits, the test instrument will indicate "good" or "bad" based on the results of the amplification measurement. If the amplifier input is not within the prescribed limits, the "no test" indication would override the "good - bad" indication.

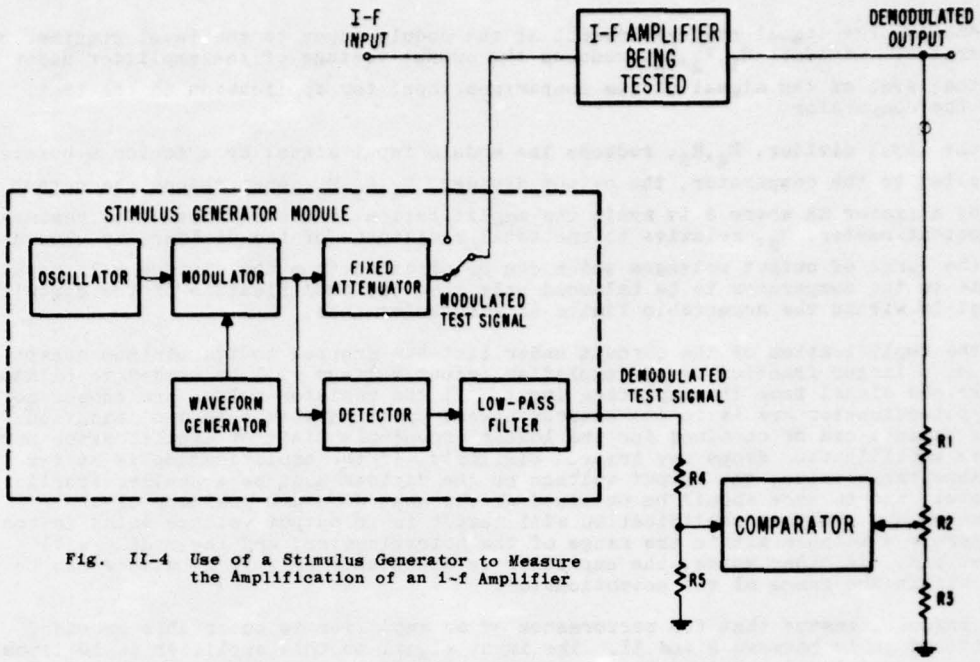



Fig. II-4 Use of a Stimulus Generator to Measure the Amplification of an i-f Amplifier

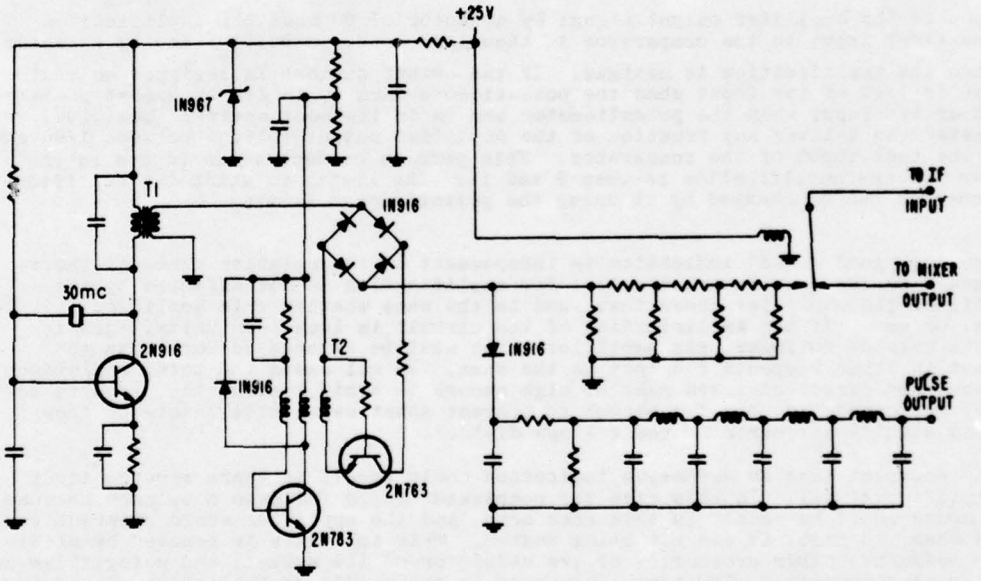



Fig. II-5 Schematic diagram of the Stimulus Generator used to permit Measurement of the characteristics of an i-f Amplifier

In many cases the maximum amplification is unimportant; the amplifier is acceptable provided its amplification is greater than a specified minimum value. In this case the amplification test can be made single-ended by eliminating $R_3$ and grounding the lower end of the potentiometer. The resistance values are chosen so that the maximum output of the potentiometer is just sufficient to equalize the inputs to the comparator when the amplification of the circuit under test is at its minimum acceptable value. The amplifier will then test "good" only when its amplification is equal to or higher than this value.

Even though there is no upper limit to the acceptable amplification, it is sometimes desirable to make both a high and a low limit amplification test if additional information about other properties of the circuit can be obtained by this means. For example, an increase in amplification which is caused by changes in component values may be accompanied by a decrease in bandwidth. Such is the case, for instance, if the plate load resistor of an RC-coupled amplifier increases beyond its design limits (initial tolerance plus permissible drifts caused by environmental changes or aging). An amplification test with both an upper and lower limit would detect this change in bandwidth through the resulting change in amplification.

2. Inserting a test signal:

In the amplification test previously described, the amplifier under test is performing its normal function, and the normal input signal is used to provide the signals for the fault location test. In some cases this may not be practicable, for example:

a. When the characteristics of the normal signal make it undesirable or impossible to use it, e.g. when the signal is intermittent, as in a receiver.

b. When the input signal level is lower than the minimum signal required by the test instrument.

In these cases, a stimulus generator can be used. A stimulus generator is a circuit built into the prime equipment solely for the purpose of providing a test signal. In the normal operation of the prime equipment it is not used; it is energized only when the test instrument is connected to those test sockets which require a signal or signals from the stimulus generator for test purposes. It is packaged as a replaceable module which may be tested and replaced, if defective, in the same way as any other module in the equipment.

For the testing of sine wave amplifiers, the stimulus generator might be an oscillator which provides a signal having the amplitude and frequency required by the circuit under test. The signal could be modulated, if necessary. For testing pulse amplifiers, a stimulus generator which produces pulses having the desired properties would be required. In some cases such a generator might be used for testing RC-coupled amplifiers, even though the normal input to the amplifier was a sine wave. Other types of stimulus generators include sawtooth, square wave, and noise generators.

The stimulus generator may be continuously connected to the circuit under test and energized only when it is required for testing, or it may be connected only during a test sequence. Jumpers are provided on the test socket for this purpose. Switches can also be provided, if necessary, to isolate the circuit under test during the test sequence from those preceding and following it.

At most, only a few stimulus generators will be required in the testing of any electronic system. In most systems, the addition of one stimulus generator module to furnish the signal input to the system will permit the entire system to be tested by FIST methods. In Radar Set AN/SPS-46, to which FIST techniques were applied, a single stimulus generator module was added to furnish an input signal for testing the i-f strip. After feeding through the i-f strip, this same stimulus generator was the source of the test signal for all the video stages in the signal path between the output of the i-f strip and the indicator. All the other signals employed by the radar set were generated internally; these internally generated signals were used to test the remaining modules of the system.

Figure II-4 illustrates the stimulus generator used to test the radar i-f amplifier module by providing a modulated signal at the intermediate frequency. The output of the modulator section of the stimulus generator is a pulse modulated i-f signal having an amplitude of a few volts. This i-f signal is attenuated so that a normal amplitude input signal is fed to the i-f amplifier being tested. The demodulated output of the

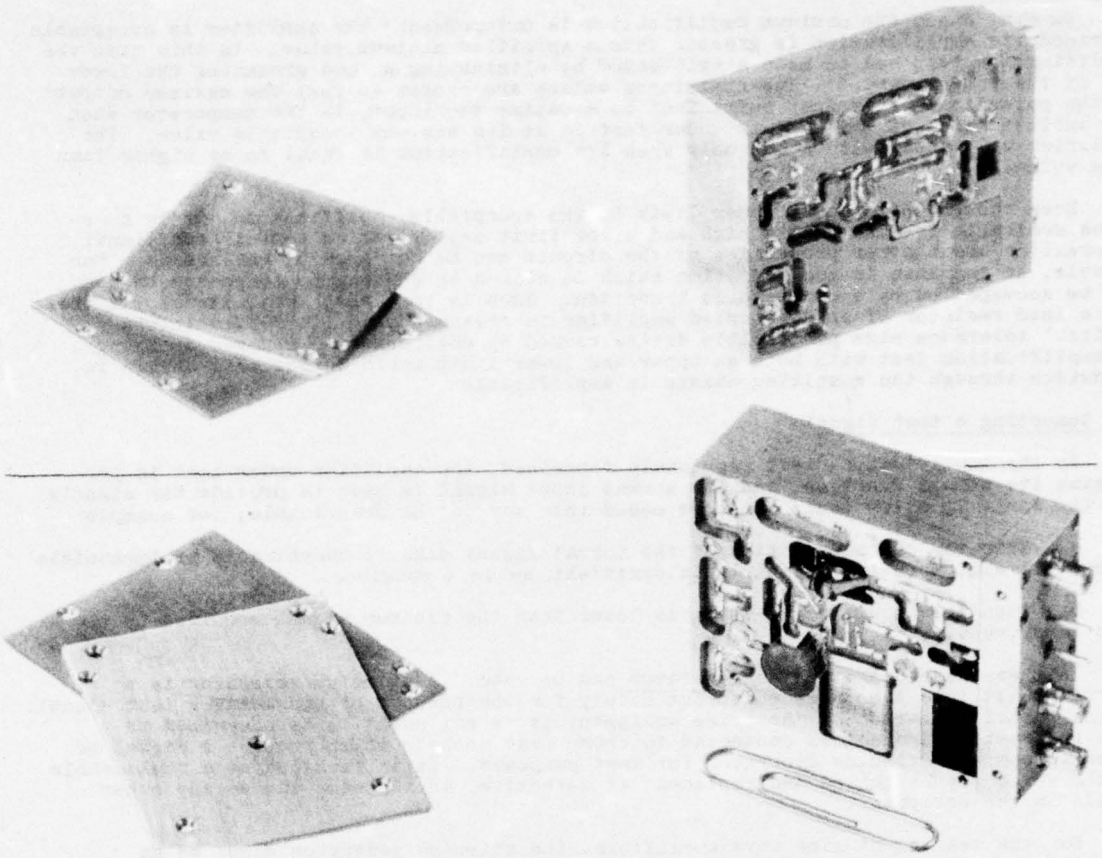


Fig. II-6 Front and Rear Views of the Stimulus Generator used to Measure the Characteristics of an i-f Strip

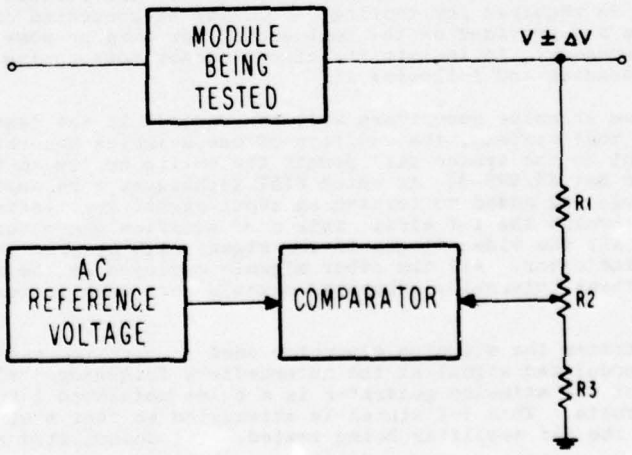



Fig. II-7 Measurement of peak-to-peak Voltage

module under test is connected to the test input of the comparator by means of divider $R_1, R_2, R_3$. The range of potentiometer $R_2$ is selected to permit a balance to be obtained within the normal limits of the amplification of the module under test.

Within the stimulus generator, another detector connected to the output of the modulator provides a demodulated test signal through a low pass filter to the comparison input of the comparator. In some cases this demodulated signal may be connected directly to the comparator; Figure II-4 shows the general case in which a voltage divider, $R_4, R_5$, is used to provide the proper level signal for the comparison input of the comparator. If the resistor values in the two dividers are correctly chosen, the comparator will indicate a balance only when the amplification of the module under test is within its specified limits.

Since both inputs to the comparator are a function of the amplitude of the signal produced by the stimulus generator, reasonable variations in this amplitude will not affect the balance indication at the output of the comparator. The output of the comparator is therefore an indication of the amplification of the circuit under test. The output level of the stimulus generator can be checked prior to testing the i-f amplifier to assure that the amplifier under test will operate within its range of linear amplification during the measurement.

The stimulus generator module itself can be checked by measuring its pertinent characteristics at the demodulated-test-signal terminal; this checks all the parts of the module except the fixed attenuator. There is no practicable way of checking the attenuator, since the level of the modulated test signal of the stimulus generator is much lower than the minimum signal level required by the comparator. The fixed attenuator is one of the more reliable components in the stimulus generator, however, so that the reliability of the test will not be seriously impaired by this omission.

The schematic diagram of this stimulus generator is shown in Figure II-5, and the actual module in Figure II-6. This is one of the more complex stimulus generators since it generates a modulated test signal, yet it demonstrates the practicability of building such a stimulus generator in a small physical size using standard components.

## VOLTAGE AMPLITUDE

A voltage may be measured by comparing its amplitude with a reference. The voltage is applied to the test input of the comparator, and the reference signal to the comparison input as shown in Figure II-7. The voltage being measured is reduced to a level within the range of the comparator by voltage divider, $R_1, R_2, R_3$, the resistance values of which are chosen so that a balance will be obtained only when the voltage being checked is within its specified limits.

Both AC and DC voltages can be measured, but since the comparator is basically a device for measuring the peak-to-peak amplitude of AC voltages, a chopper must be used to convert a DC voltage to an AC voltage before it is applied to the comparator. Provisions are made for this in the test instrument. AC voltages can be measured in the presence of DC by blocking the DC with a capacitor. The DC component of a signal which contains both AC and DC components can be measured by adding a low pass filter to the transformation network.

Pulse amplitudes are measured in the same way as AC voltages, but special precautions may be required when measuring the amplitude of a pulse or square wave in which overshoot or droop is present. Since the comparator is sensitive to changes in peak-to-peak amplitude, it may respond to changes in the overshoot or droop unless the transformation networks are designed to prevent this. The extra components required should be added to the transformation network only when changes in overshoot or droop are likely to cause unacceptable errors in the measurement of amplitude.

Figure II-8 illustrates three pulses with the same amplitude but which will be interpreted as different amplitudes by the peak-to-peak detectors of the comparator. The comparator would correctly indicate the amplitude of pulse a, but would indicate amplitude plus overshoot of b, and amplitude plus twice the overshoot of c. If the overshoot of b and c are relatively constant, a transformation network consisting of a simple voltage divider can be used to measure amplitude provided the resistors in the divider are selected to allow for the overshoots.

If changes in the overshoot are likely to obscure changes in amplitude, however, the overshoot must be eliminated before the amplitude can be measured. This can be done by adding a low pass filter section to the transformation network as illustrated in Figure II-8. If the time constant of the filter, $R_4$,$C_1$, is chosen to eliminate the overshoot, a waveform which is of the same amplitude as the original pulse but without overshoot will be applied to the comparator.

As shown in Figure II-9, the presence of droop in a rectangular waveform also increases the peak-to-peak amplitude relative to the pulse amplitude. If the droop is constant or changes by only a small amount relative to the changes in amplitude that are to be measured, a simple voltage divider will suffice as a transformation network and the presence of droop can be allowed for in the selection of the resistor values. If droop must be eliminated before the amplitude can be measured, the low frequency components of the signal can be removed by a high pass filter. The divider resistors are then chosen to reduce the resulting signal, which is twice the amplitude of the original but independent of droop, to a level that can be compared with the reference.

## OVERSHOOT

A technique for the measurement of overshoot is illustrated in Figure II-10. The waveform being measured is partially integrated to obtain a waveform without overshoot for the comparison input, while a waveform containing the overshoot is applied to the test input. The divider resistors are chosen to allow for the difference in the peak-to-peak amplitudes of the two waveforms, and the range of the potentiometer is selected to permit adjustment of the output of the test divider to obtain a balance for any overshoot within the prescribed percentage limits.

When overshoot must be measured, it is usually because it is an undesired property of the pulse, which is tolerated provided it does not exceed a specified maximum percentage of the pulse amplitude. In the transformation network for the measurement of overshoot, divider, $R_1$,$R_2$,$R_3$, must be designed so that its output can be adjusted to the level of the comparison input for any acceptable amount of overshoot from zero to the maximum specified. Maximum output will be required from the divider when there is no overshoot on the pulse being measured. For this condition the comparator will indicate a balance when $R_1/(R_2 + R_3) = R_4/R_5$.

Minimum output is required from potentiometer $R_2$ when the overshoot reaches its maximum acceptable value. If n is the maximum permissible overshoot expressed as a fraction of the pulse amplitude, and $R_1$ is made equal to $R_4$, $R_3$ must equal $R_5/(1 + n)$, and $R_2 = nR_3$.[1] (It is not necessary to make $R_1$ equal to $R_4$. If $R_1$ is a multiple of $R_4$, the specified values for $R_2$ and $R_3$ must be multiplied by the same factor.) The value of $C_1$ must be chosen to eliminate the overshoot but preserve the amplitude of the waveform being measured.

For these resistor values, the outputs of the two dividers can be made equal, and the comparator will therefore indicate a balance provided the overshoot is within the specified limits. Overshoot is thus measured as a percentage of the amplitude of the waveform. The "no test" feature of the test instrument must be used to preclude the possibility of obtaining a "good" indication when there is no output from the module under test. (See last paragraph, page II-5.)

## DROOP

Droop is also measured as a percentage of the signal amplitude. In this case the comparison input is obtained by differentiating the waveform under test to obtain a waveform whose amplitude is independent of the droop. The signal applied to the test input is obtained from a simple voltage divider and contains the same percentage droop as the waveform being measured.

The resistors used in the comparison divider, $R_4$,$R_5$, (Figure II-11) are chosen to obtain the attenuation or isolation required. Since the differentiated signal at

[1] $R_2 = nR_5/(1 + n)$ may be used if more convenient.

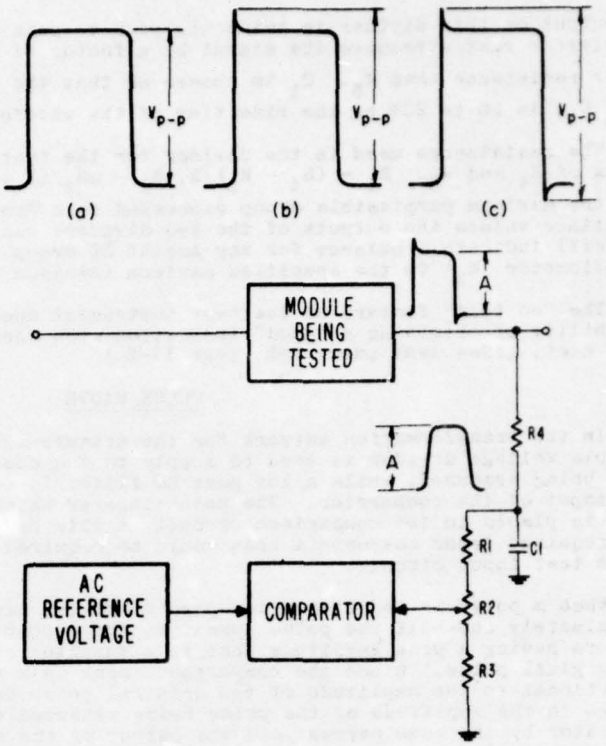



Fig. II-8 Measurement of Pulse Amplitude when Overshoot is Present

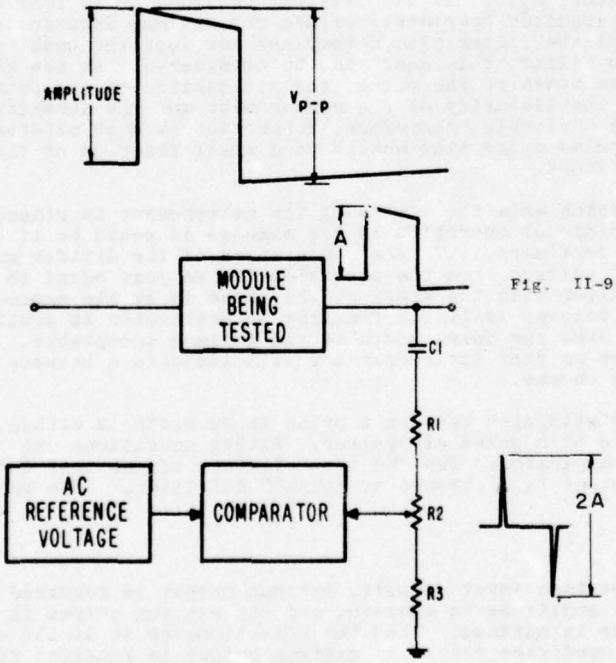



Fig. II-9 Measurement of Pulse Amplitude when Droop is Present

the output of this divider is twice the peak-to-peak amplitude of the original signal, the divider must attenuate the signal by a factor of more than two, i.e., $R_4$ must be a higher resistance than $R_5$. $C_1$ is chosen so that the time constant of the divider, $R_4$,$R_5$,$C_1$, is 20 to 25% of the rise time of the waveform under test.

The resistances used in the divider for the test input are determined from the values of $R_4$ and $R_5$. $R_1 = (R_4 - R_5)/2$, $R_2 = nR_5/(1 + n)$,[2] and $R_3 = R_5/(1 + n)$, where n is the maximum permissible droop expressed as a fraction of the amplitude. For these resistance values the outputs of the two dividers can be made equal so that the comparator will indicate a balance for any amount of droop from zero (maximum output from potentiometer, $R_2$) to the specified maximum (minimum output from the potentiometer).

The "no test" feature of the test instrument must again be used to preclude the possibility of obtaining a "good" indication when there is no output from the module under test. (See last paragraph, page II-5.)

## PULSE WIDTH

In the transformation network for the measurement of pulse width, Figure II-12, a simple voltage divider is used to supply to the comparison input a replica of the pulse being measured, while a low pass RC filter is used to transmit a signal for the test input of the comparator. The potentiometer which sets the limits of the measurements is placed in the comparison circuit in this case because this circuit configuration required fewer components than would be required if the potentiometer were placed in the test input circuit.

When a pulse is fed into a low-pass RC filter having a time constant greater than approximately one-half the pulse duration, the output of the filter will be a triangular waveform having a peak amplitude that is a function of both the width and amplitude of the original pulse. Since the comparison input is a pulse whose amplitude is also proportional to the amplitude of the original pulse but independent of its width, changes in the amplitude of the pulse being measured will change both inputs of the comparator by the same percent and the output of the comparator will be a function of pulse width only.

The time constant of the filter, $R_4$,$C_1$, is not critical provided it is long enough to preclude the charging of the capacitor completely within the maximum duration of the pulse. Since the time constant of the filter also determines how much the peak amplitude of the pulse is attenuated by the filter, this must also be considered. As the time constant increases relative to the width of the pulse, the attenuation of the peak amplitude increases, but so does the linearity of the measurement and the sensitivity to changes in pulse width. It is advisable, therefore, to provide as much attenuation as possible in the filter. The pulse decay time should be a small fraction of the pulse duration for good measurement accuracy.

Although the potentiometer which sets the limits of the measurement is placed in the comparison circuit, the principle of operation is the same as it would be if the potentiometer were placed in the test circuit.[3] The resistances of the divider are chosen so that the highest output voltage from the potentiometer is just equal to the voltage output of the low-pass filter when the width of the pulse is at its maximum acceptable value, and the lowest voltage available from the potentiometer is equal to the output voltage of the filter when the pulse width is the minimum acceptable. The potentiometer can then be adjusted so that the comparator will indicate a balance for any pulse width within the limits chosen.

The test instrument cannot distinguish between a pulse whose width is within acceptable limits, and the absence of a pulse altogether. Either conditions can produce identical inputs to the comparator. The "no test" feature of the test instrument must therefore be used to detect the "absence of signal" condition. (See page II-5.)

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[2] or $R_2 = nR_3$.

[3] When the potentiometer is in the test input circuit, maximum output is required from the potentiometer when the signal amplitude is minimum, and the minimum output is required when the signal amplitude is maximum. When the potentiometer is in the comparison input circuit, the opposite condition prevails; maximum output is required from the potentiometer when the signal amplitude is maximum, and vice versa.

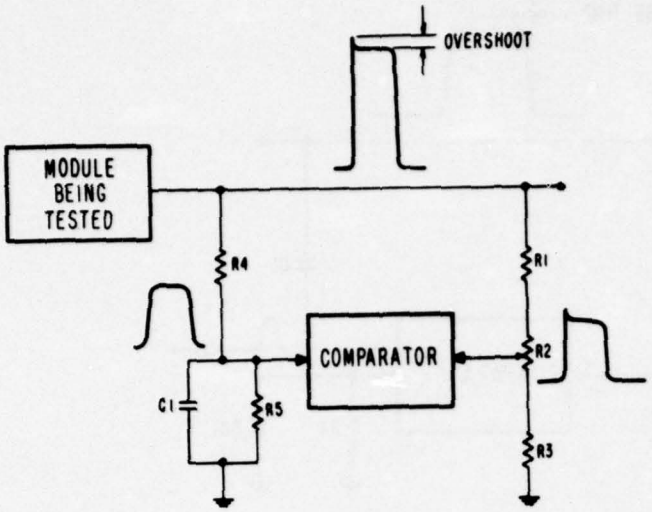


Fig. II-10 Measurement of Overshoot

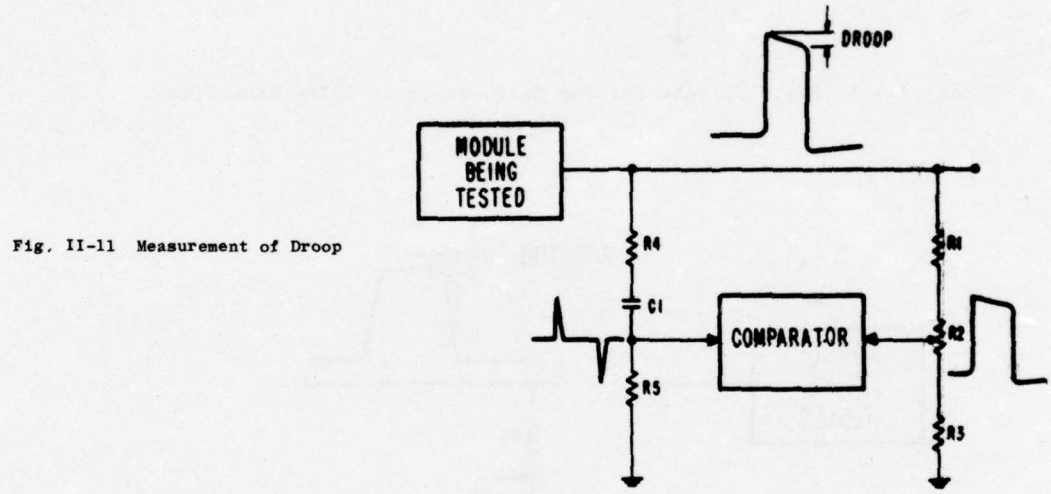


Fig. II-11 Measurement of Droop

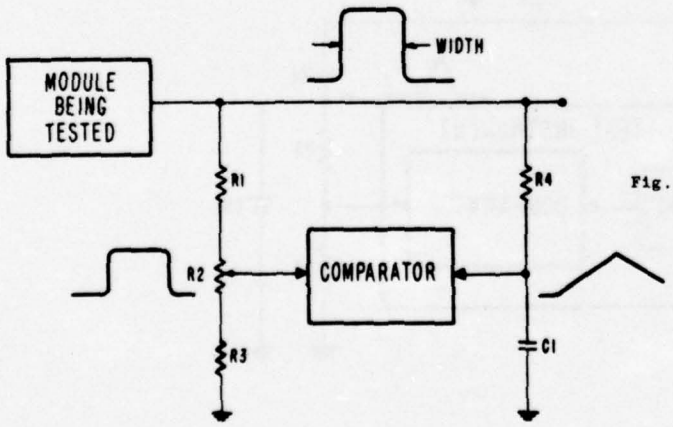


Fig. II-12 Measurement of Pulse Width

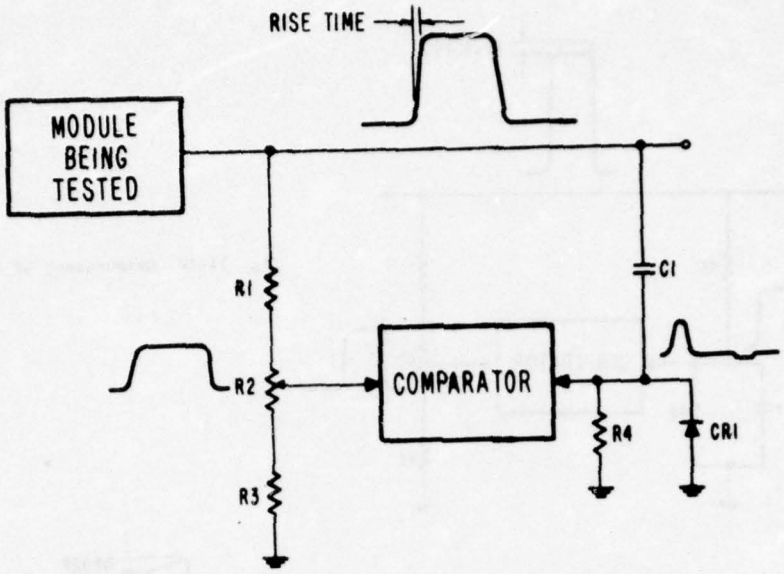


Fig. II-13 Basic Circuit for the Measurement of Pulse Rise Time

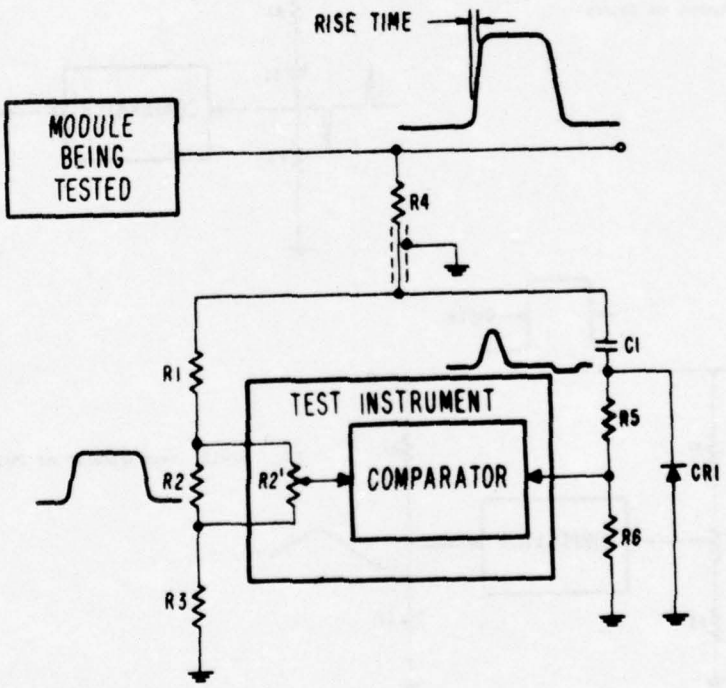


Fig. II-14 Practical Circuit for the Measurement of Pulse Rise or Decay Time

## PULSE RISE OR DECAY TIME

The test signal for measuring pulse rise or decay time is obtained from a high-pass filter which connects the pulse source to the test input of the comparator, Figure II-13. The output of this filter is a signal whose amplitude is proportional to the amplitude and transition times of the original pulse. It is compared with a signal whose amplitude is proportional to the amplitude of the pulse being measured, but independent of its transition times.

When a pulse is fed into a high-pass RC filter, the output of the filter will be a pair of pulses of opposite polarity, one occurring at the leading and one at the trailing edge of the original pulse. The peak amplitude of the first pulse is a function of the amplitude of the original pulse and the transition time of its leading edge (rise time), while the peak amplitude of the second pulse is a function of the amplitude of the original pulse and the transition time of its trailing edge (decay time). Since the two pulses are of opposite polarity, either can be selected by clipping the other by means of a suitably polarized diode. Thus either the rise or the decay time can be selected for measurement.

The filter used to measure rise or decay time is designed for maximum sensitivity to small percentage changes in transition time by making the time constant of the filter equal to the time constant of the transition time being measured. (See Appendix A.) Under these conditions the peak output of the filter will be about 52% of the original pulse amplitude, and the percentage change in the amplitude of the filter output for a given percentage change in transition time will be maximum.

Attenuator, $R_1$,$R_2$,$R_3$, furnishes a signal to the comparison input of the comparator whose amplitude is proportional to the amplitude of the original signal. (The potentiometer is placed at the comparison input because this circuit configuration uses fewer components.) The high-pass filter, $R_4$,$C_1$, furnishes the test input to the comparator, and diode $CR_1$ clips the negative portion of the filter output. The test input is therefore a positive pulse whose amplitude is a function of the positive transition time of the signal being measured.

The time constant of the filter is made equal to the longest (worst) acceptable transition time of the signal being measured. Faster transition times will cause the filter output amplitude to increase, while slower changes will cause a decrease in the amplitude. The resistors of the comparison divider are then chosen to provide a minimum output from the potentiometer which is 52% of the amplitude of the pulse being measured, and a maximum output which is equal to the peak amplitude of the filter output when the transition time is minimum. The potentiometer can then be adjusted to achieve a balance only if the transition time is within the specified limits. Since the amplitudes of both the comparison and test inputs to the comparator vary directly with the amplitude of the pulse being measured, the output of the comparator is substantially independent of changes in pulse amplitude.

In many applications it is only necessary to measure transition time to a single limit; the transition time must not be longer than a specified maximum value, but no minimum transition time is specified. In this case $R_1$ may be omitted, so that the potentiometer output can be varied from 52% to 100% of the pulse amplitude.

The circuit of Figure II-13 is not usually suited to practical applications because in most cases the output of the high-pass filter would be higher than the level required by the comparator, and because capacitor $C_1$ would load the circuit under test when diode $CR_1$ conducts. Furthermore, the output impedance of the module under test may not be suitable for driving the high-pass filter which requires a low impedance driving source for proper operation. A divider can be used to provide the necessary isolation from the filter and at the same time provide a low-impedance driving source, but cannot be used to reduce the signal amplitude to a level within the input range of the comparator because the level so obtained would be too low for the proper operation of the diode.

A practical transformation network for the measurement of transition time contains three dividers as illustrated in Figure II-14. Isolation resistor, $R_4$, and the comparison input divider, $R_1$,$R_2$,$R_3$, together comprise the divider which provides isolation for the diode. Resistor, $R_4$, is usually installed at the module socket and connected

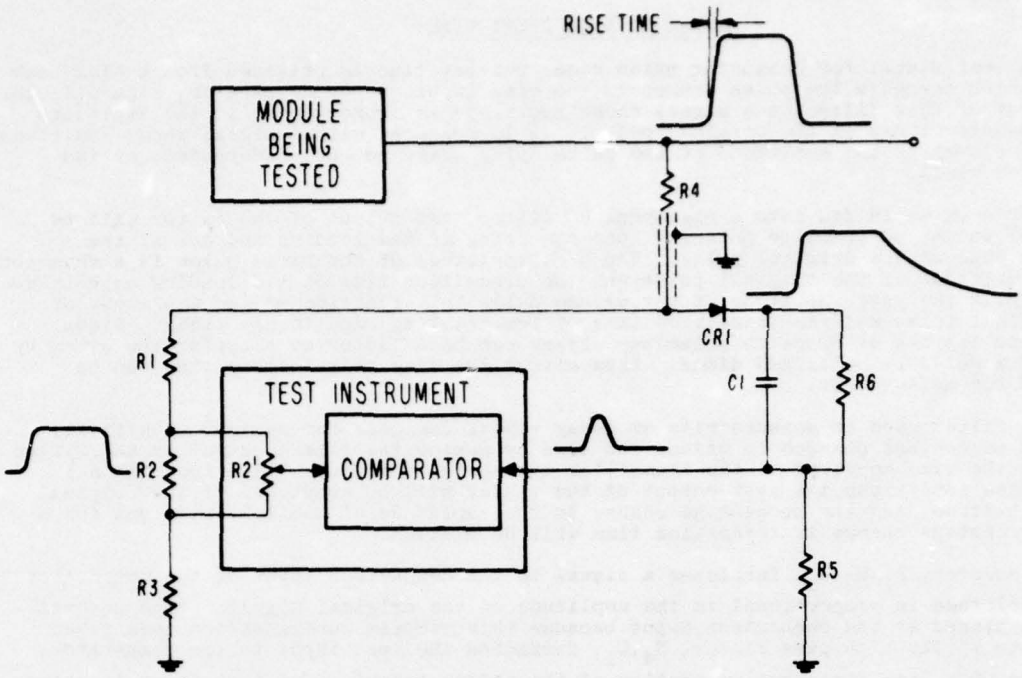


Fig. II-15 Alternate Circuit for Measuring Rise Time

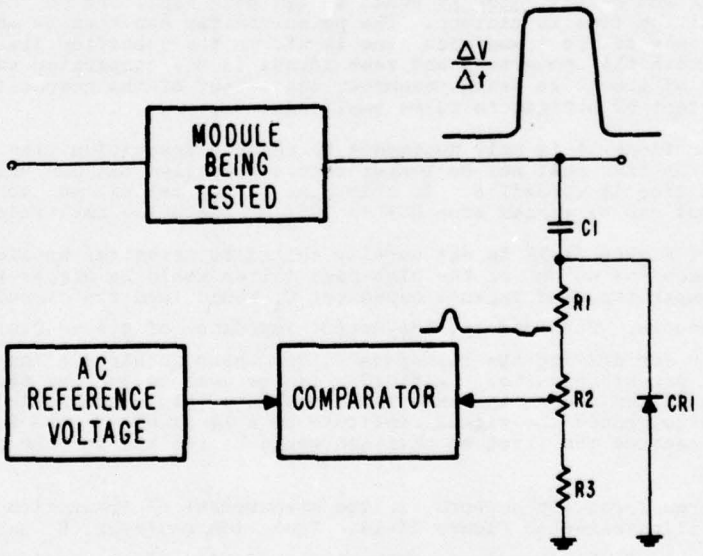


Fig. II-16 Basic Circuit for the Measurement of Rate of Rise

to the test socket by a shielded lead to provide isolation between the module under test and the test socket. Divider, $R_5$,$R_6$, reduces the signal across the diode to a level that is within the range of the comparator, and the comparison divider, $R_1$,$R_2$,$R_3$, provides the output range needed to permit a balance to be obtained at any transition time within the range specified. (Potentiometer, $R_2$, could now be moved to the test input circuit, if desired, without requiring any increase in the number of components. As previously mentioned, resistor $R_1$ may be omitted in the case of a single limit.)[4]

For the measurement of rise time, the alternate circuit shown in Figure II-15 can be used to improve the attenuation of the unwanted pip at the trailing edge. In this circuit the diode, $CR_1$, is placed in series with the input to the differentiating network. For the duration of the leading edge of the pulse the diode conducts, permitting capacitor $C_1$ to charge through resistor $R_5$. Since the time constant $R_5C_1$ is chosen to emphasize changes in rise time, the output of the differentiating network during the leading edge of the pulse is identical with the one previously described. At the conclusion of the pulse the diode ceases to conduct, and the capacitor discharges through $R_6$. The amplitude of the pip at the trailing edge of the pulse is minimized by making the time constant $R_6C_1$ as long as possible consistent with the requirement that $C_1$ be completely discharged before the beginning of the next pulse. This circuit cannot be used for the measurement of decay time since the development of diode back-bias to clip the unwanted pip requires that capacitor, $C_1$, be charged when the corresponding transition occurs.

In any of these transformation networks, the disappearance of the output signal completely would result in a "good" indication from the comparator since both of its inputs would be equal (zero). To prevent the module from testing good when it is not, one of the other properties of the module or of its stimuli must be used as a qualifying measurement. (See last paragraph, page II-5.)

If the signal whose transition time is being measured is less than about a volt in amplitude, diodes cannot be used effectively to attenuate the unwanted pip. Measurement of the transition time must then be based on the amplitude of both pips in the output of the high-pass filter. This can be done when changes in the performance of the module affect both transition times, and therefore the amplitudes of both pips, equally, or when performance changes result in a change in only one of the transition times. When module performance changes sometimes affect one of the transition times and sometimes both, the measurement cannot be based on the amplitude of both pips unless large errors can be tolerated. When rise and decay times are significantly different, as is often the case, the time constant of the high-pass filter may be selected for its sensitivity to changes in the faster of the two transition times. (See Appendix A, Figure A2.)

## RATE OF RISE OR DECAY

In some applications the rate of rise (or decay) is a more important criterion of performance than rise (or decay) time. This is true of many waveforms used as triggers; for example, the rate of rise of the input to a parallel-triggered blocking oscillator is more important than its rise time. The transformation network used for the measurement of rate of rise or decay differs from that used for rise time, in the time constant of the high-pass filter and the comparison voltage. The discussion which follows is based on measurement of rate of rise, but applies equally to measurement of rate of decay unless otherwise noted.

A typical transformation network for the measurement of pulse rate of rise is illustrated in Figure II-16. If the time constant of the high-pass filter is made much smaller than the transition time of the signal being measured, the output of the filter will approach the true derivative of the input signal. In practice, a time constant

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[4] If the potentiometer is placed in the test input circuit, maximum output from the potentiometer is required when the transition time is maximum (minimum test signal amplitude) rather than when transition time is minimum as is the case when the potentiometer is in the comparison circuit, and minimum potentiometer output is required for the shortest transition time when the filter output is greatest. In this case if there is no minimum limit to the transition time it is the grounded resistor of the divider, $R_3$, that should be omitted.

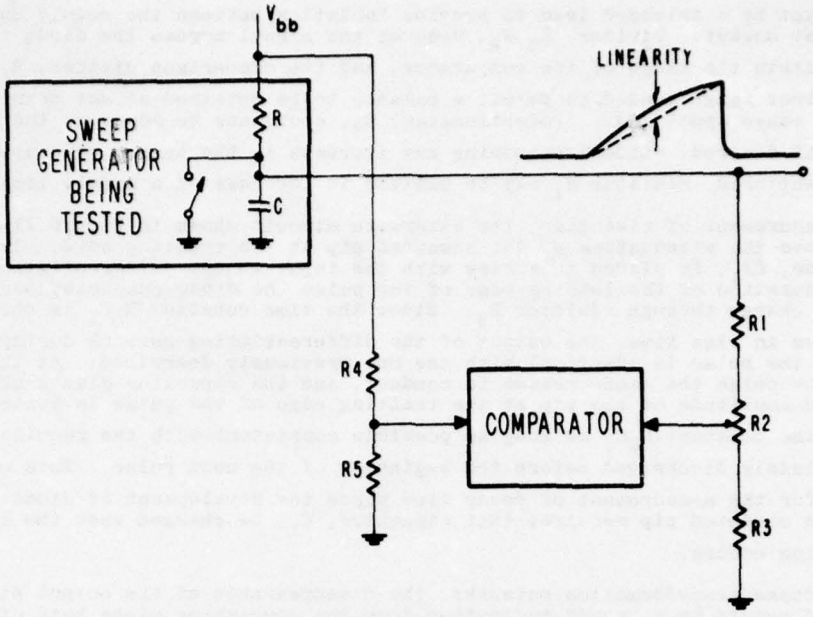


Fig. II-17 Measurement of Linearity

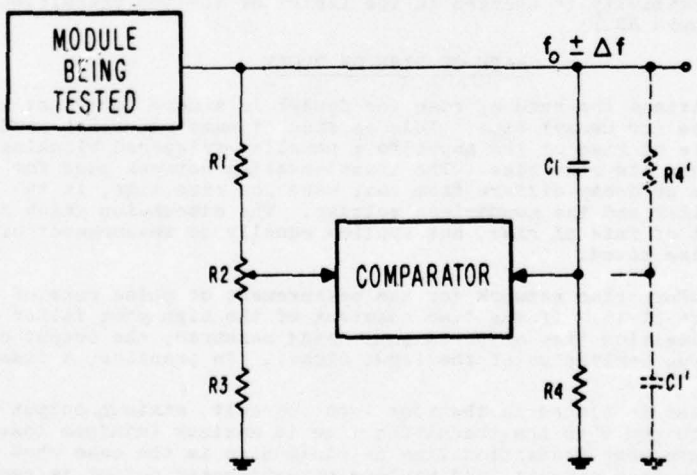


Fig. II-18 Simple Transformation Network for the Measurement of Low Frequency

0.05 or less of the transition time of the signal is satisfactory. (See Appendix B.) The values of the divider resistors are chosen so that the minimum output from the potentiometer is equal to the reference voltage when the rate of rise is maximum, and the maximum output from the potentiometer balances the reference voltage when the rate of rise is minimum. $R_3$ may be eliminated when rate of rise is tested to a minimum limit only, since in this case the maximum acceptable rate of rise is infinite and the minimum output required of the potentiometer is zero. When measuring rate of rise, the method illustrated in Figure II-15 may be used.

The attenuation of the filter may be determined approximately by noting that the ratio of the output-to-input voltage of the filter is approximately equal to the ratio of the time constant of the filter to the time constant of the transition time being measured.[5] For example, if the ratio of the time constants is 0.1, the ratio of output-to-input voltage is 0.08, and if the ratio of the time constants is 0.01, the output-to-input voltage ratio is also about 0.01.

A qualifying measurement must be made in conjunction with the rate of rise (or decay) measurement to distinguish between a failure within the module and a malfunction caused by an incorrect input to the module. (See last paragraph, page II-5.)

## SWEEP LINEARITY

A simple type of sweep linearity measurement can be made by comparing the peak amplitude of the sweep against a reference. The module being tested in Figure II-17 is a simple type of sweep generator in which the switch is opened at $t = 0$ and closed at $t = t_1$ (the switch is actually either a vacuum tube or transistor which is either off or in a saturated condition). The sweep output voltage, $V_o$, is the voltage developed across the capacitor during this interval.

The sweep is a segment of an exponential whose linearity depends on what fraction of the exponential is used. The linearity is a function of $t_1/RC$, while the final (peak) output voltage is a function of $t_1/RC$ and the supply voltage, $V_{bb}$; therefore the amplitude of the peak voltage is a measure of the linearity, provided the effects of supply voltage changes are accounted for. Once the limits of $t_1$, R, and C are established, the acceptable limits of the final output voltage can be calculated. This peak voltage can be compared with a reference to ascertain whether the linearity is within acceptable limits.

If the supply voltage, $V_{bb}$, is well regulated, the internal reference in the test instrument can be used. If the supply voltage is not regulated, the reference can be made proportional to the supply voltage by deriving it from the supply voltage as shown in Figure II-17. The duration of the sweep should be checked separately as a qualification measurement (see page II-5 ) to prevent changes in both the supply voltage and $t_1$ from giving erroneous results. In the phantastron circuit, time is the variable rather than voltage, so that it would be necessary to use voltage as the qualifying measurement and time as the performance measurement.

## FREQUENCY

In the transformation network for the measurement of frequency, either or both of the inputs to the comparator contain frequency-sensitive elements to convert frequency changes to voltage changes. The type of frequency-sensitive element is determined by the frequency to be measured.

For frequencies in the audio or ultrasonic range, one of the simplest transformation networks consists of an RC circuit to provide the input to one side of the comparator, and a simple resistive divider to provide input to the other, as indicated in Figure II-18. Either a high-pass section ($R_4$,$C_1$) or a low-pass section ($R_4'$, $C_1'$) can

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[5] The rise time of a pulse is usually taken to be the time required for the pulse to rise from 10% to 90% of its amplitude, and the decay time as the time required for it to fall from 90% to 10%. For convenience, this rise or decay is assumed to be an exponential in which case the time required for the change is 2.2 times the time constant of the exponential. The ratio of filter time constant to transition time constant is then $2.2RC/t_t$, where $t_t$ is the transition time.

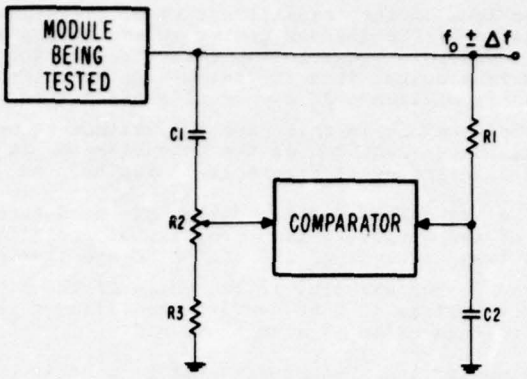


Fig. II-19 More Sensitive Transformation Network for the Measurement of Low Frequency

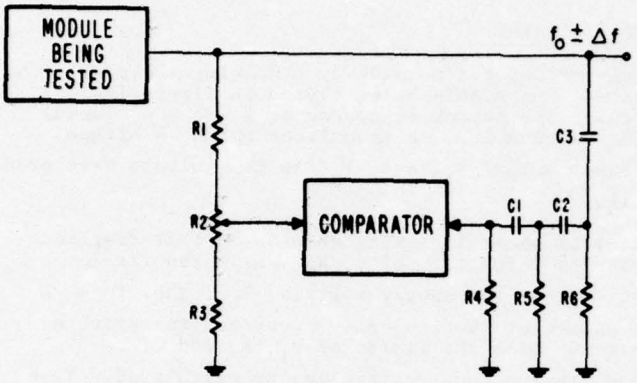


Fig. II-20 The Sensitivity of the Frequency Measurement can be Improved by using several Filter Sections

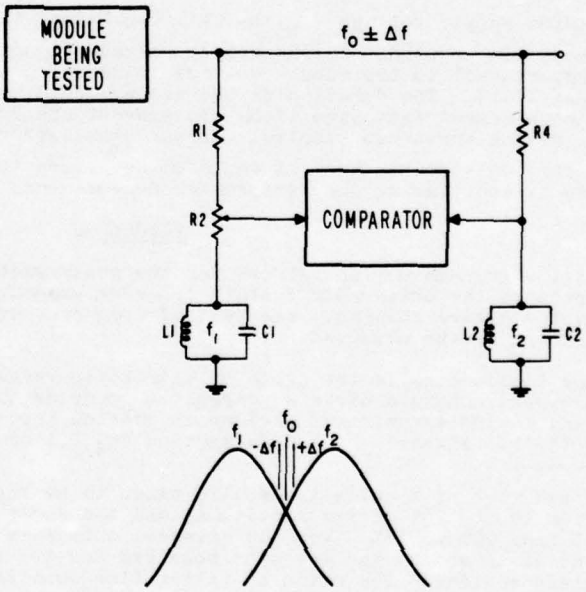


Fig. II-21 Transformation Network for the Measurement of High Frequency

be used; in either case, the time constant is chosen so that the frequency to be measured lies outside the pass band in the region where the amplitude response is falling off. The output amplitude of the RC section is then a function of frequency, while the output of the resistive divider is not. Changes in frequency will therefore affect the equality of the two inputs to the comparator, while changes in the amplitude of the signal being measured will not, since amplitude changes affect both inputs to the comparator proportionally.

In selecting values for $R_4$ and $C_1$ (or $R_4'$ and $C_1'$), both the attenuation between the input and output of the network and the sensitivity of the output voltage of the network to changes in frequency must be considered. If the sensitivity of the network is defined as the ratio of the percentage change in output voltage to the percentage change in input frequency, plots of sensitivity versus attenuation would be identical for both the high- and low-pass sections, except for the direction of the change. They would show a rapid increase in sensitivity to about 0.5 (percentage change in output voltage is one-half the percentage change in frequency) as the attenuation increases from 0 to 3 db and then a more gradual increase to a sensitivity of 1 as the attenuation increases from 3 to 20 db, remaining at 1 for attenuation greater than 20 db.

For a given attenuation, therefore, either a high- or a low-pass RC section in the transformation network will yield the same sensitivity. The component values required will not be the same, however. If the high-pass section is used, the time constant, $R_4C_1$, must be greater than the period of the frequency being measured, while for the low-pass section the time constant, $R_4' C_1'$, must be less than the period. The component values used in the high-pass section will therefore be larger than those used for a low-pass section of the same sensitivity and attenuation. In either case, the component values are determined by the amount of attenuation required.

Greater sensitivity for the same amount of attenuation can be obtained by using two RC networks as shown in the transformation network of Figure II-19. For any given attenuation this transformation network will have double the sensitivity of the network shown in Figure II-18. For attenuation of 20 db or more, the percentage change in the voltage seen by the comparator will be double the percentage change in the frequency of the signal being measured. This is usually a more effective way of increasing sensitivity than cascading two sections on one side of the transformation network.

The sensitivity can be further increased by increasing the number of RC sections as shown in Figure II-20. The increase in sensitivity is proportional to the number of RC sections used if the attenuation is very high (40 db or more), while little is gained by increasing the number of sections if the attenuation required is less than 6 db. For attenuations between 6 and 40 db the additional sections will improve the sensitivity but not in proportion to the number of sections added. For example, at an attenuation of 26 db the sensitivity of the network of Figure II-20 is 2.5 times greater than the sensitivity of a single section. Best sensitivity is obtained if the additional sections do not appreciably load the preceding ones. The sensitivity can be still further increased by replacing the resistive divider of Figure II-20 with a low-pass RC section or sections.

The transformation network illustrated in Figure II-21 is used to measure frequencies in the medium- and high-frequency range. In this transformation network each input to the comparator is obtained from a divider which is composed of resistance in series with a parallel-resonant circuit. As indicated by the diagram, which shows the voltage amplitude at each input to the comparator as a function of frequency, one of the resonant circuits is tuned to a frequency higher than the frequency being measured, and the other to a lower frequency.

The resistance values are chosen relative to the parallel-resonant resistance of the LC circuits so that the comparator will indicate a balance with the potentiometer in its mid-position when the frequency is at its nominal value, $f_o$. If the frequency changes, the voltages across the tuned circuits will change in opposite directions and must be compensated by a change in the setting of the potentiometer to obtain a balance on the comparator. The potentiometer resistance is chosen so that balance can be restored within the range of the potentiometer only if the frequency change is within acceptable limits.

In this network the sensitivity of the output of the network to changes in frequency is a function of the Q of the tuned circuits and the separation between their anti-

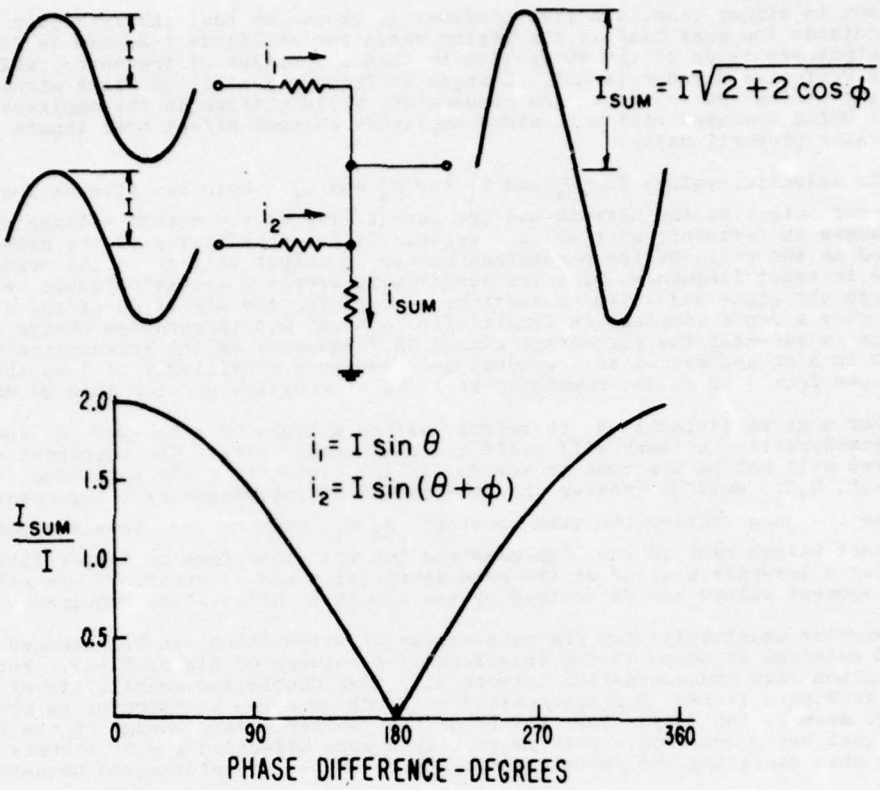


Fig. II-22 Peak Amplitude of the Sum of Two Sine Waves of Equal Amplitude as a Function of the Phase Difference Between Them

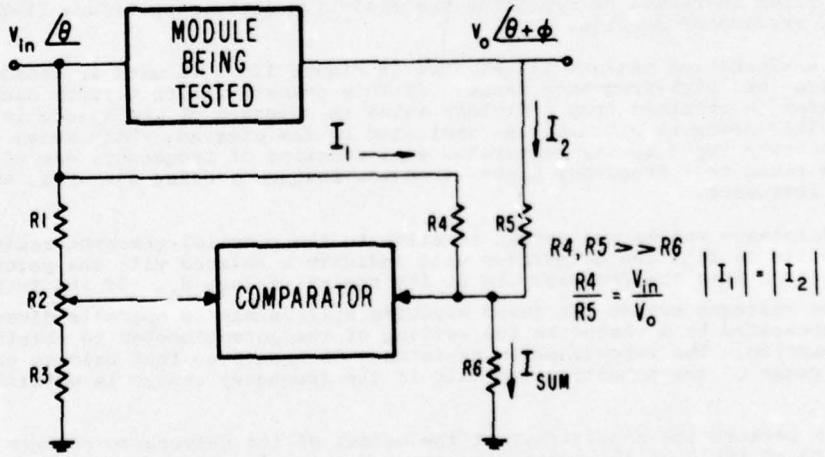


Fig. II-23 Transformation Network for the Measurement of Phase Shift

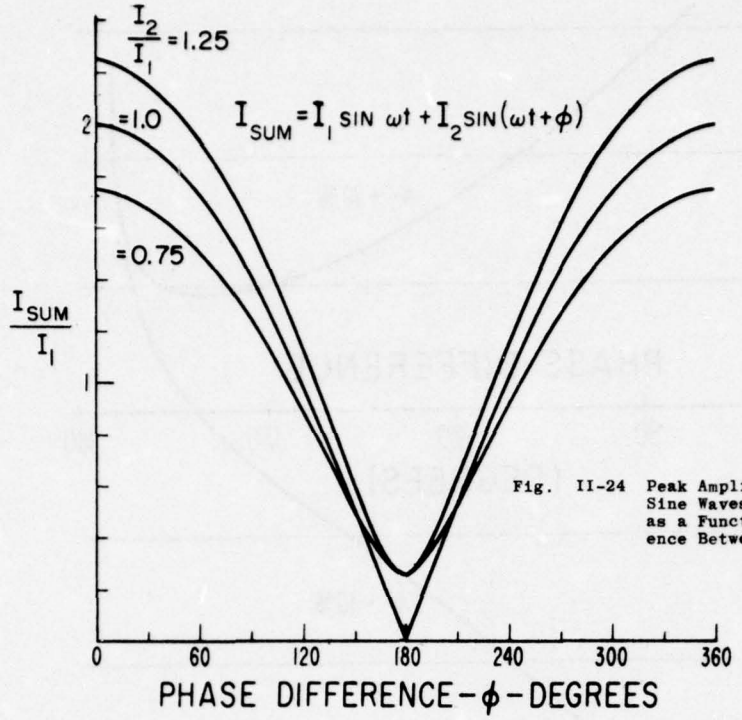


Fig. II-24 Peak Amplitude of the Sum of Two Sine Waves of Unequal Amplitude as a Function of the Phase Difference Between Them.

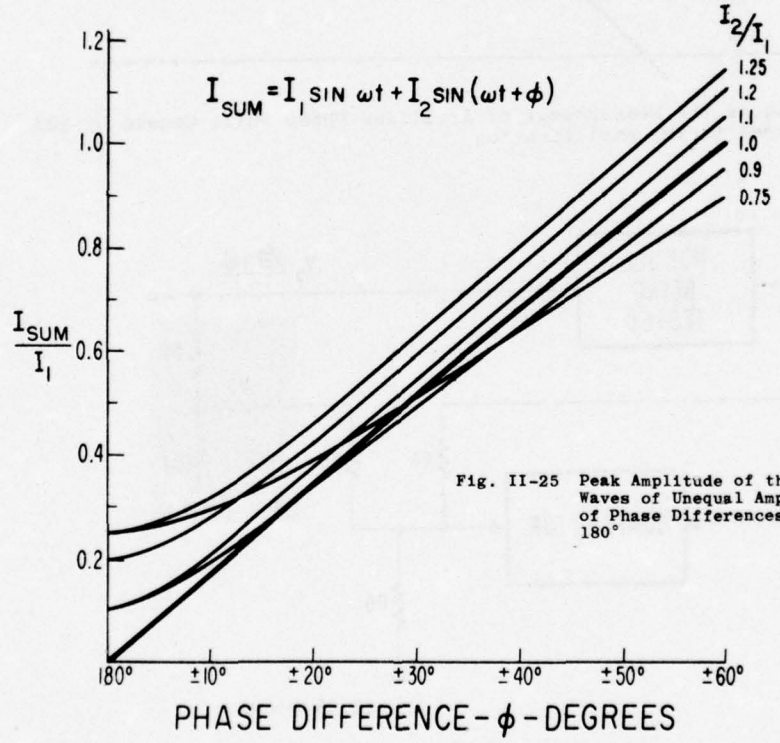


Fig. II-25 Peak Amplitude of the Sum of Two Sine Waves of Unequal Amplitude as a Function of Phase Differences in the Vicinity of 180°

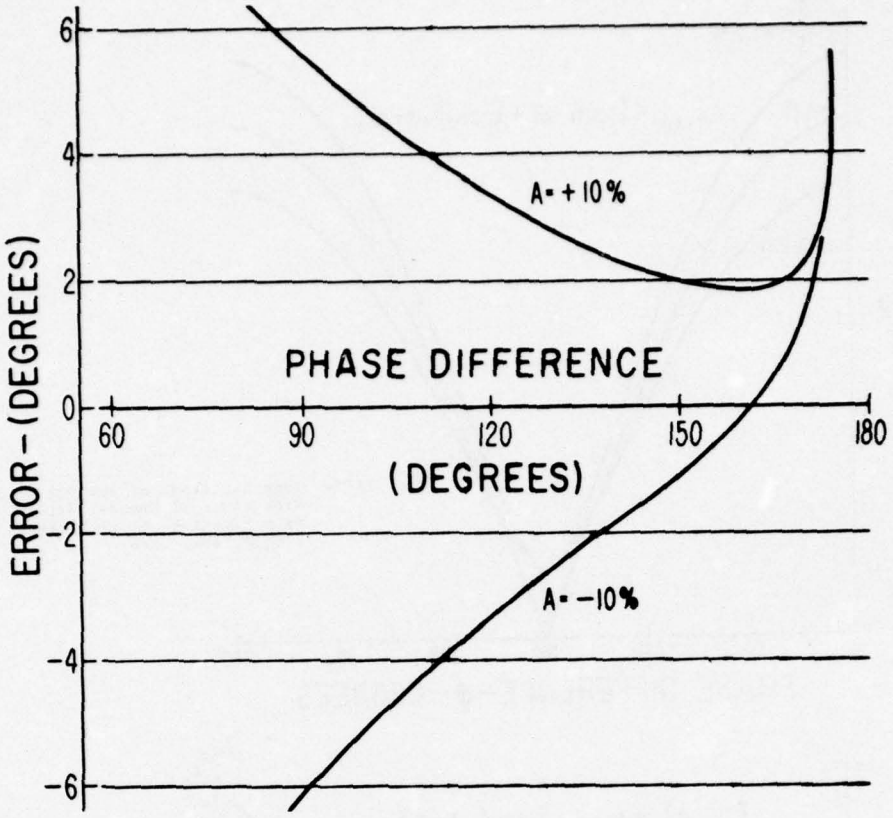


Fig. II-26 Error in the Measurement of Amplifier Phase Shift Caused by 10% Changes in the Amplification

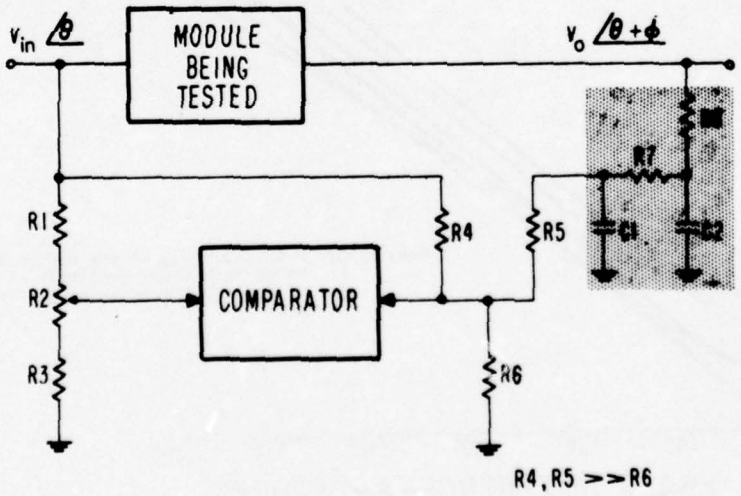


Fig. II-27 Transformation Network used to Minimize Errors in the Measurement of Amplifier Phase Shift Caused by Changes in Amplification

resonant frequencies. The sensitivity will be greatest when the resonant circuits are tuned to frequencies which are separated by approximately 1.4 times their band-width.

## PHASE SHIFT

The transformation network for the measurement of phase shift is based on the fact that the sum of two sine waves of the same amplitude and frequency is a sine wave whose peak amplitude is proportional to the phase difference between the two. The relationship between the amplitude of the sum and the amplitude of either input is shown in Figure II-22. In the vicinity of zero degrees, the amplitude of the sum is relatively insensitive to changes in phase shift, but within the 180° ± 60° region the sensitivity is good, and the relationship between the amplitude of the sum and the phase difference between the inputs is almost linear. However, it should be noted that at 180° there is no voltage available.

A simple transformation network employing this principle for the measurement of phase shift is illustrated in Figure II-23. It is assumed that the phase difference between the output and input signals of the module being tested is within the second quandrant but not so close to 180° that insufficient signal output is available.

Resistors, $R_4$, $R_5$, and $R_6$ form a simple summing network. If the voltage across $R_6$ is to be proportional to the sum of $V_{in}$ and $V_o$, $R_4$, and $R_5$ must be so large compared with $R_6$ that the effect of $R_6$ on the current in either of the input branches is negligible. For best indication of phase shift, the resistances of $R_4$ and $R_5$ must also be chosen so that the currents flowing through them are equal even though $V_{in}$ and $V_o$ are not. Under these conditions, the voltage across $R_6$, which is applied to one input of the comparator, is proportional to the phase difference between the input and output voltages of the module under test. The reference input to the comparator is obtained by attenuating the input voltage of the module being tested by means of divider, $R_1, R_2, R_3$.

The ratio between the output voltage of the summing network and the input voltage to the module under test is the product of the ratio of $R_6$ to $R_4$ and a factor which depends on the phase difference between the input and output voltages of the module. This factor, plotted in Figure II-22, permits the output amplitude of the summing network to be determined once the permissible phase shift limits are known. After these voltage limits have been determined, the values for the resistances of divider, $R_1, R_2, R_3$, can be calculated so that the output of potentiometer, $R_2$, can be varied within the same limits. The output of the comparator will then indicate a balance only when the phase difference between $V_{in}$ and $V_o$ is within the acceptable limits.

Because both inputs to the comparator are a function of the signal level at the input to the module under test, changes in this level will not affect the phase shift measurement. The accuracy of the measurement is affected by changes in the amplification of the module, however; this unbalances the currents into the two inputs of the summing network by the same percent that the amplification changes.

The sum of two sine waves of unequal amplitude is plotted as a function of the phase difference between them in Figures II-24, and II-25. Figure II-24 is plotted over 360° to show the general form of the curves, while Figure II-25 expands the region in the vicinity of 180°. The curve for equal input signals, $I_2/I_1 = 1$, is also shown for comparison. When the two input signals are equal, their sum is an almost linear function of the phase difference between them from 120° to 180° and from 180° to 240°. When the two input signals are not equal, their sum is an insensitive indicator of changes in phase shift within 10 to 20 degrees on either side of 180°, but thereafter their sum becomes an almost linear function of phase shift for the next 40 to 50°.

The problem in checking the phase shift of an amplifier, however, is that the amplification varies within specified limits, with the result that the ratio between the current inputs to the summing network is not constant. Since the calibration of the reference divider, $R_1, R_2, R_3$, must be based on a particular ratio of the inputs to the summing network, the changes in amplification will introduce errors in the phase shift measurement.

TABLE II-2

Sum of Two Sine Waves of Unequal Amplitude Which Differ in Phase by 205°

| Ratio of inputs[1] $I_2/I_1$ | Output[1] $I_{sum}/I_1$ |
|---|---|
| 1.25 | 0.54 |
| 1.2 | 0.50 |
| 1.1 | 0.46 |
| 1.0 | 0.43 |
| 0.9 | 0.42 |
| 0.8 | 0.44 |
| 0.75 | 0.45 |

[1] Input current $I_1$ is taken as a reference, and both $I_2$ and the sum are given relative to $I_1$.

TABLE II-3

Errors Caused by Changes in the Ratio of the Input Currents to a Summing Network. $I_1$ is Taken as the Reference Current.

| | (a) Transformation Network Designed for Equal Input Currents, $I_2 = I_1$ | | (b) Transformation Network Designed for $I_2 = 1.1I_1$ | |
|---|---|---|---|---|
| Ratio of Inputs | Maximum Phase* Difference | Error | Maximum Phase* Difference | Error |
| 1.25 | 198° | -7° | 200° | -5° |
| 1.2 | 200.5° | -4.5° | 202.5° | -2.5° |
| 1.1 | 203° | -2° | 205° | 0° |
| 1.0 | 205° | 0° | 207° | +2° |
| 0.9 | 205.5° | +0.5° | 207.5° | +2.5° |
| 0.8 | 204° | -1° | 206.5° | +1.5° |
| 0.75 | 203.5° | -1.5° | 206° | +1° |

*At this phase difference the potentiometer will be at the upper limit of its range. This value will coincide with the 205° design limit only at the ratio of input currents for which the transformation network was designed.

For example, consider an amplifier whose amplification may vary ± 25% and whose phase shift is acceptable provided it does not exceed 205° (180° + 25°). The construction of the amplifier is such that the phase shift will always be in excess of 185°. Since for this example it is only necessary to measure to the upper limit, 205°, resistor $R_3$ can be omitted from reference divider, $R_1$,$R_2$,$R_3$. At 205° the output of the summing network will lie between $0.42V_{in}R_6/R_4$ and $0.54V_{in}R_6/R_4$,[6] depending on the amplification at the time the measurement is made. Since changes in amplification cause the same percentage change in the ratio of input currents, Figure II-25 can be used to termine the extend to which the output of the summing network varies for a given phase shift as the amplification changes. This has been done for a phase shift of 205° and the results tabulated in Table II-2. This information can be used to determine the maximum voltage required across potentiometer, $R_2$. The best value will vary with the application, and is chosen to minimize consequences of the errors in the phase measurement caused by the amplification changes.

If the transformation network is designed for zero error at the nominal amplification of the amplifier, the maximum output of the potentiometer will be adjusted to $0.43V_{in}R_6/R_4$ and the error caused by changes in the amplification is indicated by Table II-3(a). The error is within the range +0.5°, -2° if the amplification is between 0.75 and 1.1 of its nominal value. At an amplification of 1.25 times nominal, the error rises to -7°.

The maximum error can be reduced by designing the transformation network for zero error at an amplification of 1.1 nominal. The maximum output voltage from the potentiometer would then be set at $0.46V_{in}R_6/R_4$, and the errors for changes in amplification are indicated in Table II-3(b). The error for amplification between 0.75 and 1.2 nominal has increased to ± 2.5° but the maximum error at an amplification 1.25 times nominal has been reduced to -5°. The transformation network may be designed for zero error at any other value of amplification within the prescribed range.

The errors resulting from a ± 10% change in amplification are plotted as a function of phase difference in Figure II-26. The data for the curves were obtained by designing the transformation network for equal input currents to the summing network, and determining the errors introduced when the amplification change caused a 10% difference in input currents. Within the region from 150° to 168° the error caused by the 10% change in amplification is less than 2°; by designing for a current ratio slightly greater than unity this region of minimum error can be extended to phase differences between 140° to 170°. At 120° the error is about 3.5° while for phase differences above 174° no measurements can be made. Curves plotted for other amplification ratios would be similar in shape but would differ in the magnitude of the error.

The transformation network shown in Figure II-27 employs RC sections to deliberately introduce phase shift within the transformation network. If the phase limit to be measured is in a region where the error is large, such a transformation network can be used to shift the phase so that the summing network sees a phase difference which is within the region of minimum error. A second useful application of this transformation network is for the case in which both limits of the phase difference must be measured and one is in the third quadrant and the other in the fourth. The transformation network of Figure II-27 can be used since the phase shift within the transformation network can be used to increase or decrease the total phase shift so that both limits lie in the same quadrant.

A qualifying measurement must be made in conjunction with the phase shift measurement illustrated in Figure II-23 or II-27 to distinguish between failures within the module and a malfunctioning caused by an incorrect input to the module. (See page II-5.)

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6 From Figure II-23, $V_{sum} = I_{sum}R_6$, and $V_{in} \approx I_1R_4$, so that

$$\frac{V_{sum}}{V_{in}} = \frac{I_{sum}R_6}{I_1R_4} = \frac{I_{sum}}{I_1}\left(\frac{R_6}{R_4}\right)$$

The value for $I_{sum}/I_1$ is obtained from Figure II-25.

## RESISTANCE, INDUCTANCE AND CAPACITANCE

Although testing to the piece-part level is not advocated, it can be done by the method illustrated for the measurement of resistance in Figure II-28. During the measurement, the component being tested can be disconnected from the rest of the circuit, if required, by microswitches that can be mounted on the rear of the test socket. (See last paragraph, page I-6.)

The input voltage may be furnished by the internal reference voltage of the test instrument or by a voltage obtained from the equipment being tested. The transformation network together with the resistor being tested forms two voltage dividers and the comparator is used to compare their outputs. The ratio of the resistance of $R_2$ to the total resistance of divider $R_1, R_2, R_3$ is chosen so that equality will be obtained only if the resistance being tested is within the prescribed limits.

This method will probably find most frequent applications in the measurement of physical properties rather than in the measurement of resistance. Instead of being a component of the prime equipment, the resistor may be a transducer which is placed in the module to measure light, temperature, pressure, etc. The change in resistance then becomes the measure of the physical property.

Capacitance or inductance can be measured by a similar arrangement if the transformation network is modified by replacing $R_3$ with a reactive element of the same type as the component being measured. The transformation network must be powered by a sinusoidal signal when reactance is being measured.

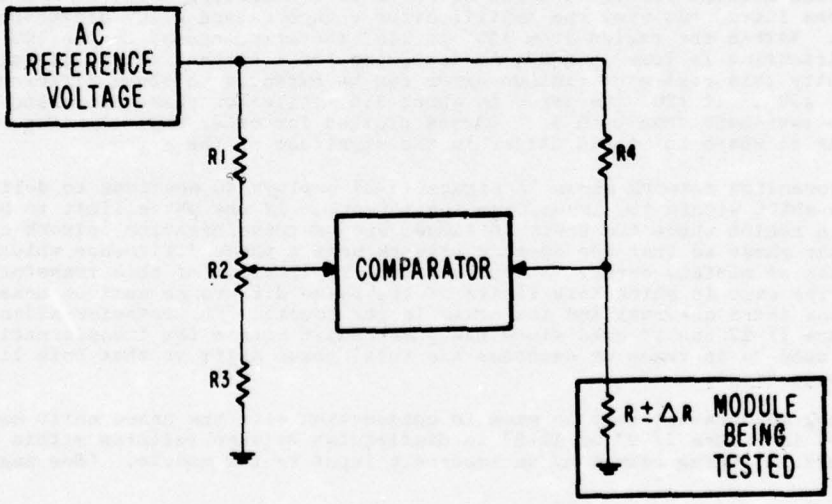


Fig. II-28 Measurement of Resistance

## APPENDIX A

### Optimum RC-filter Time Constant for Measurement of Pulse Rise or Decay Time

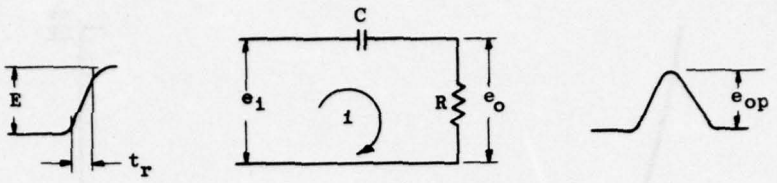


Fig. II-A1

A high-pass RC filter with its input and output waveforms is shown in Figure II-A1. It is necessary to determine the RC time-constant of the filter which will maximize the change in the peak amplitude of the filter output for a given percentage change in rise time of the input waveform. The loop equation for the network of Figure II-A1 is:

$$e_1 = \frac{1}{C}\int i\,dt + e_o \tag{1}$$

Differentiating (1) with respect to t:

$$\frac{de_i}{dt} = \frac{i}{C} + \frac{de_o}{dt},$$

$$\frac{de_i}{dt} = \frac{e_o}{RC} + \frac{de_o}{dt}. \tag{2}$$

If the leading edge is assumed to be exponential, the input waveform may be given in terms of its rise time as:

$$e_1 = E\left(1 - \epsilon^{\frac{-2.2t}{t_r}}\right), \tag{3}$$

where E is the input pulse amplitude and $t_r$ is the rise time from 10% to 90% of the pulse amplitude. Substitution of the input waveform (3) into (2) gives

$$\frac{2.2E}{t_r}\epsilon^{\frac{-2.2t}{t_r}} = \frac{e_o}{RC} + \frac{de_o}{dt}. \tag{4}$$

Equation (4) is a linear differential equation in $e_o$. Solving for $e_o$ gives

$$e_o = \left(\frac{-E}{t_r - \frac{t_r}{2.2RC}}\right)\epsilon^{\frac{-2.2t}{t_r}} + K\epsilon^{\frac{-t}{RC}} \tag{5}$$

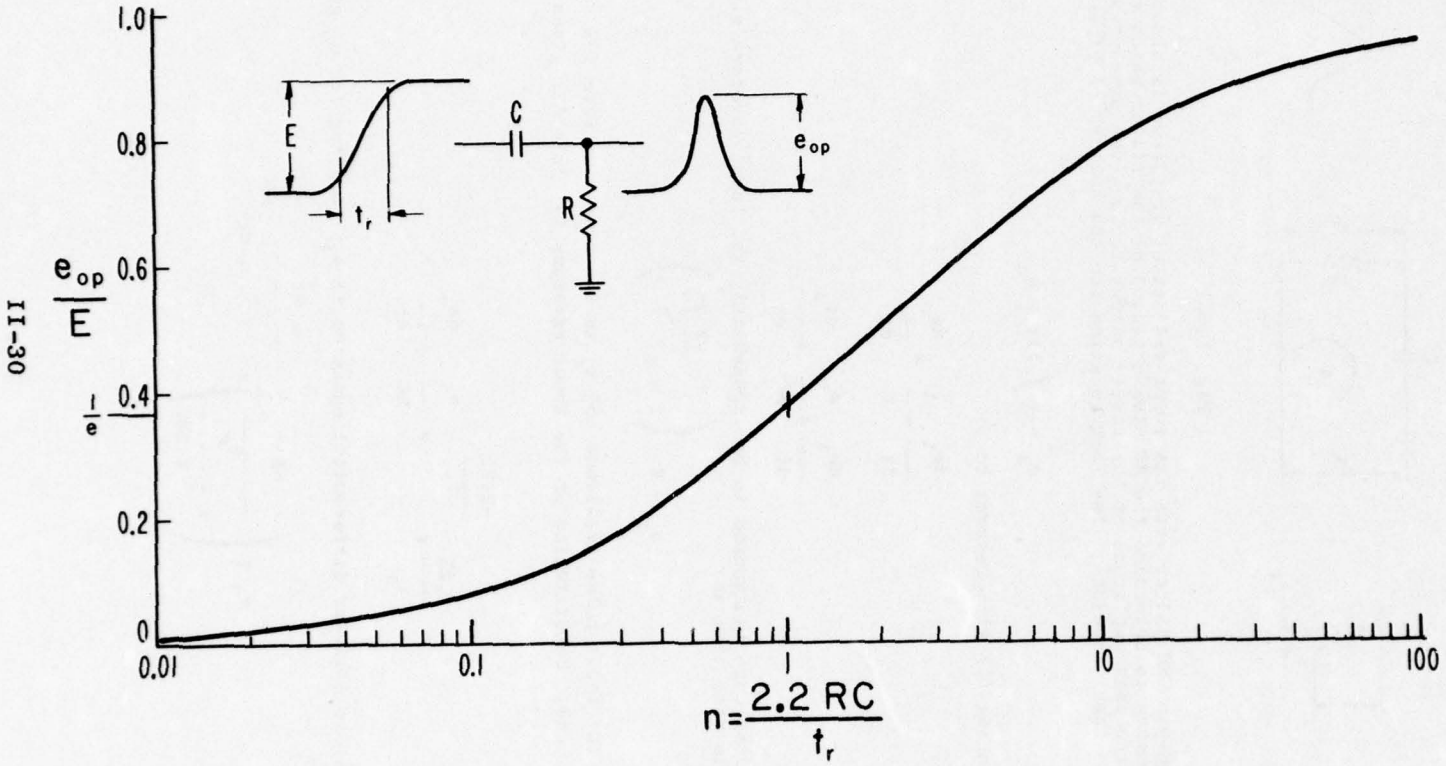


Fig. II-A2 Effect of RC Time Constant on the Peak Amplitude of the Output Voltage

The complete solution of (4) is determined by the condition that at t = 0 the voltage across the resistor is 0. The final solution then becomes

$$e_o = \frac{E}{1 - \frac{t_r}{2.2RC}} \left[ \epsilon^{\frac{-t}{RC}} - \epsilon^{\frac{-2.2t}{t_r}} \right]. \qquad (6)$$

Since an expression for the peak or maximum value of $e_o$ is desired, equation (6) must be differentiated with respect to t and equated to 0. Equation (6) is first simplified by substituting $n = \frac{2.2RC}{t_r}$, where n is the ratio of the filter time constant to the waveform time constant. Equation (6) then becomes

$$e_o = \frac{En}{n - 1} \left[ \epsilon^{\frac{-t}{RC}} - \epsilon^{\frac{-nt}{RC}} \right]. \qquad (7)$$

Differentiating (7) yields

$$\frac{de_o}{dt} = \frac{En}{n - 1} \left[ \frac{n}{RC} \epsilon^{\frac{-nt}{RC}} - \frac{1}{RC} \epsilon^{\frac{-t}{RC}} \right],$$

$$\frac{de_o}{dt} = 0 \quad \text{at} \quad t = \frac{RC \ln n}{n - 1}. \qquad (8)$$

Substituting (8) into (7) and simplifying gives the peak amplitude of $e_o$ as

$$e_{op} = En^{\frac{1}{1-n}} \qquad (9)$$

Figure II-A2 shows a plot of equation (9) on a semi-logarithmic scale. In the range n = 0.5 to n = 5, the slope is fairly constant and near its maximum. There is, however, a value of n where the slope is greatest. Taking the first derivative of (9) and expressing the slope as a function of the percentage change in n, the following expression is obtained.

$$\frac{\Delta(e_{op}/E)}{\Delta n/n} = n^{\frac{1}{1-n}} \left[ \frac{1}{1 - n} + \frac{n \ln n}{(1 - n)^2} \right]. \qquad (10)$$

Using numerical methods, the slope is found to be maximum at n = 2.2.

Since $n = \frac{2.2RC}{t_r}$ is the ratio of the filter time constant to the waveform time constant, the change in the peak amplitude of the filter output for a given percentage change in signal rise time will be greatest when n = 2.2 or $RC = t_r$. The corresponding peak value of the filter output voltage ($e_{op}$, equation (9)) is 0.518E.

The preceding derivations hold only if the filter does not appreciably load the source of the input waveform.

## APPENDIX B

### Optimum RC-filter Time Constant for Measurement of Rate of Rise or Decay

The rate of rise of a pulse is defined as the amplitude (E) divided by the rise time ($t_r$) measured between the 10 and 90 percent points. The expression for the peak output voltage (equation (9), Appendix A) is

$$e_{op} = En^{\frac{1}{1-n}} \tag{1}$$

As n approaches zero or gets very small, the expression becomes approximately

$$e_{op} \simeq En. \tag{2}$$

Since n is the ratio of the filter time constant to the input waveform time constant, $n = \frac{2.2RC}{t_r}$, the peak output voltage then becomes a linear function of the rate of rise

$$e_{op} \simeq \left[2.2RC\right]\left[\frac{E}{t_r}\right]. \tag{3}$$

Thus if a filter is designed so that its time constant is very much smaller than the time constant of the input waveform, the peak amplitude of the filter output will be a function of the rate of rise of the input waveform.

There are, however, practical considerations involved in the selection of the filter time constant. If n is made too small, the amplitude of the filter output will be too small to allow a practical measurement. When $n = 0.1$, $e_{op} = 0.079E$, and if $n = 0.01$, $e_{op} = 0.00956E$. The value of n chosen depends on how accurately a rate of rise measurement must be made, and on the amplitude of the filter output. In most cases a value of $n = 0.1$ will be satisfactory. For a constant rate of rise and a 50% change in pulse signal amplitude, the peak output amplitude of the filter will change only about 4.5% when $n = 0.1$.

# PART III

## THE DESIGN OF A FIST TEST INSTRUMENT

### INTRODUCTION

The test instrument illustrated and briefly described as a voltage comparator will be described in greater detail here. This section first discusses the objectives of the test instrument program. Following the discussion of objectives, the basic principles of a test instrument containing one comparator are described and illustrated, after which the need for multiple cells in a test instrument is established. (Each comparator is referred to as a cell of the instrument.) The remainder of the section describes the circuitry, over-all operation, and self-testing methods of a four-cell general-purpose test instrument presently under development.

### OBJECTIVES

When developing a test instrument, certain objectives must be kept in mind. The degree to which such objectives can be achieved governs to a large extent the practicality and acceptance of the FIST maintenance techniques. The following five objectives have been used as guide lines in the development of the general-purpose test instrument:

1. General-Purpose Instrument:

In order that the fault location methods may be applied to a wide variety of equipments, the test circuitry is divided between the test instrument and the transformation networks. The test instrument should be versatile enough to accept the output of a large variety of transformation networks. This permits the incorporation of fault isolation in many types of equipment, thus reducing the obsolescence problem so often encountered in test equipment designed for testing only one special equipment.

2. Simplified Interpretation:

The operation of the test instrument should be as simple and straightforward as possible and should require the manipulation of a minimum number of controls. It is assumed that the maintenance technician has a limited knowledge of test equipment and may have no knowledge of the function performed by the prime equipment being tested. Also, an operator of a piece of prime equipment might assume the task of "trouble-shooting" his equipment in an emergency if the fault location method is simple to learn and easy to apply. Thus the indication of performance should be on a good - bad basis requiring the operator to make no decisions.

It is realized that the use of good - bad indicators must be approached with caution. Good - bad lights, often disparagingly referred to as idiot lights, are frequently resented by the technician because of the obvious implication. Furthermore, skilled technicians may feel that a good - bad indication does not give them enough information to support their intuitions. However, when all factors are weighed, good - bad indicators appear to best fit the requirements for an indication means that may be interpreted quickly and accurately by a wide variety of personnel whether trained or untrained.

3. Self-Checking Capability:

The test instrument should contain certain self-checking features to permit the determination of proper functioning at any time. This objective is important to gain operator acceptance and confidence in the instrument, thereby, enabling him to make fullest use of its capabilities.

4. Instrument Simplicity:

The basic principles of the instrument should be kept as simple as possible with variations among circuits held to a minimum. The use of stable circuits which do not require frequent adjustment or calibration is also desirable. If adjustment is

necessary the circuit principles should be simple and straightforward so that rapid and trouble-free repair and calibration procedures can be employed.

5. Physical Size:

The size and weight of the test instrument should be sufficiently small so that it could be easily carried in one hand. An example of typical size and weight desired is that encountered in an ordinary VTVM. The intent is for the maintenance technician to bring the test instrument to the malfunctioning prime equipment, place the instrument on the floor or an available shelf and test on the site of the prime equipment.

## THE BASIC TEST INSTRUMENT

Beginning with the simple concepts illustrated in Figure III-1, various elements will be added or modified in sequence, to facilitate a clear understanding of the principles being used. Figure III-1 shows a basic test instrument testing an amplifier module. The elements enclosed within the heavy border constitute the test instrument and those within the dotted border comprise a cell. The instrument of Figure III-1 is termed a one-cell instrument. A cell consists of two wide-band amplifiers of identical gain, driving peak-to-peak detectors whose outputs are fed to a differential amplifier. The output of the differential amplifier is connected to a null indicator which in its simplest form is a zero-center microammeter.

The wide-band amplifiers provide the gain needed for linear detection of low-level signals. It is important only that the relative gain of the two amplifiers be the same, since the equality of two signals is being measured. A peak-to-peak detector provides a DC output voltage which is proportional to the peak-to-peak voltage of an AC signal. This allows the amplitude comparison of dissimilar waveforms.

Equal peak-to-peak voltages having like or unlike waveforms when connected to the inputs of the wide-band amplifiers, will produce a zero reading on the null indicator.

Briefly reviewing how a module performance limit measurement is accomplished, it should be recalled that if a null can be obtained at some setting of $R_2'$, the assembly is considered good. However, the purpose is not to find a nulling point with $R_2'$ but to determine whether the null indicator passes through null when the arm of potentiometer $R_2'$ is moved from one extreme to the other. A single-pole double-throw switch can be substituted for $R_2'$ and the effect will be the same. This is illustrated in Figure III-2. An indication of acceptable performance is obtained when the null meter alternately swings through zero as the switch $S_1$ is alternately switched across $R_2$. Although a zero-center microammeter is perhaps the simplest type of null indicator, from an operator's point of view, a meter swinging back and forth is an unconventional indication. In addition, the sampling rate of switch $S_1$ is limited by the inertia of the meter movement.

Since it is necessary only to determine whether the output of the differential amplifier passes through zero or changes polarity, the indicator meter has been replaced by a zero crossing detector shown in Figure III-3. The output of the zero crossing detector drives either a good (green light) or a bad (red light) indicator. As switch $S_1$ connects the input of one side of the comparator alternately to the ends of $R_2$, the zero crossing detector will excite the good indicator if the output of the differential amplifier passes through zero, or the bad indicator if it does not pass through zero. Switch $S_1$ can be either an electrically driven switch actuated at a low sampling rate or a hand operated push button switch. This report will discuss a test set using manually actuated switches.

The section on transformation techniques also described the need for a comparison reference voltage when measuring voltage amplitudes. When a voltage amplitude measurement is made, the reference voltage is tied to one input of the test instrument by connecting together points 2 and 3 of Figure III-4 by means of a jumper on the test socket of the equipment. The method of programming tests by means of jumpers on the test sockets will be covered in greater detail when the over-all operation of the breadboard model test instrument is described.

Because of the many problems associated with direct-coupled amplifier circuits, a practical method of amplifying low frequencies and DC signals is to convert them to a

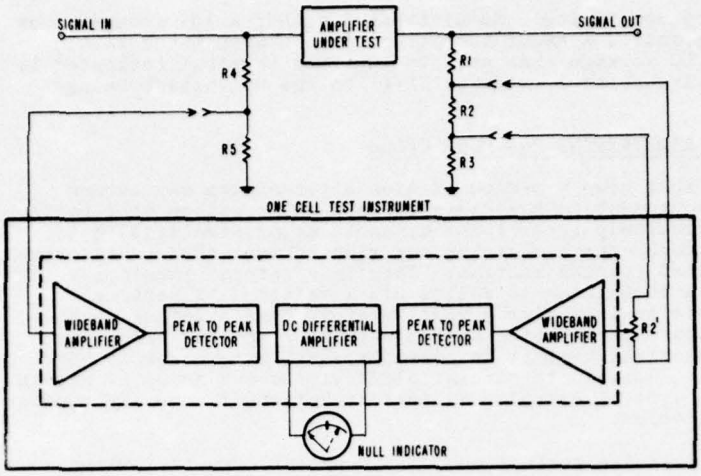


Fig. III-1 Basic One-Cell Test Instrument Amplification Measurement

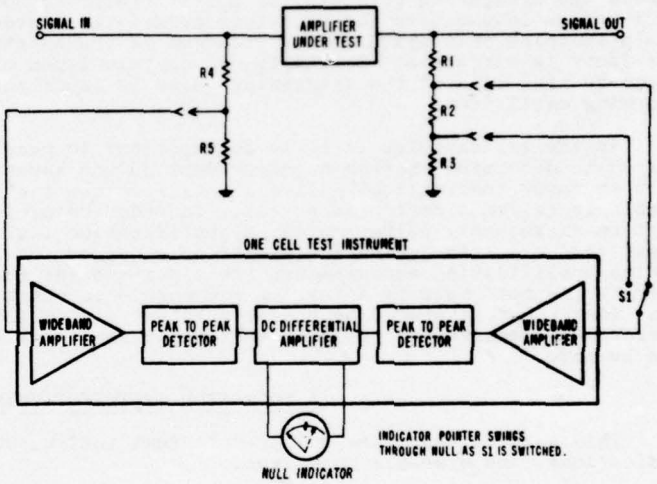


Fig. III-2 Basic One-Cell Test Using Switched Input

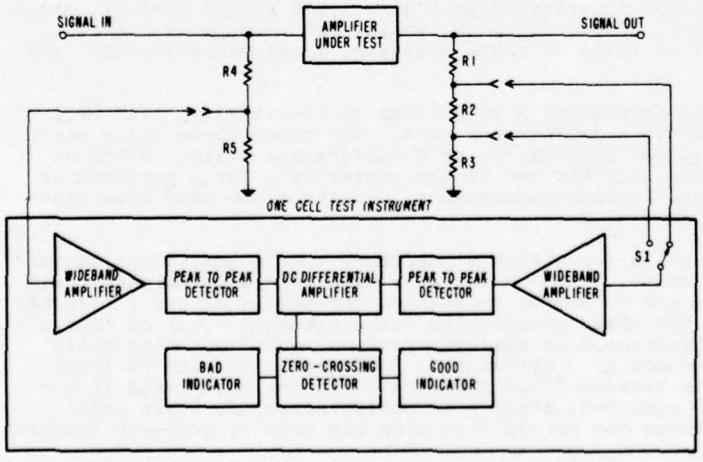


Fig. III-3 One-Cell Test Instrument With Zero Crossing Detector Replacing Null Indicator

higher frequency AC signal before amplifying. Modulation of either a low frequency or a DC signal is accomplished by means of a shunt chopper contained within the test instrument (Figure III-4). Any DC voltage that must be measured is first converted to a pulsating DC signal by connecting point 1 (Figure III-4) to the DC voltage being measured.

## THE REQUIREMENT FOR MORE CELLS

It is reasonable to expect that over a period of time a technician may become familiar enough with the prime equipment to associate certain malfunctions with particular test sockets. This is particularly true if the assembly associated with that socket is a frequent offender. Hence the technician may wish to test that test socket first before following a prescribed testing routine. This is a natural reaction which must be satisfied by allowing the technician to follow his intuition. A serious consequence of forcing the technician to follow a routine every time a failure occurs will be his negative attitude toward the testing system resulting in less effective maintenance. Therefore it is important that it be possible for him to check in any random order he chooses. However, certain technicial difficulties can arise if random order testing is practiced. Two typical examples of test situations frequently encountered will illustrate the basic problem.

Suppose it is necessary to test the performance of a monostable multivibrator module in which the important characteristic is the output pulse amplitude. The limits of the output pulse amplitude can be tested in the manner previously described (by connecting the reference voltage to the comparison input of the comparator). However, should the triggering pulse to the multivibrator be absent or degraded, the multivibrator will become inoperative and no output pulse will appear. The test instrument, however, would indicate that a failure had occurred in the multivibrator module although actually the fault is elsewhere. Similarly, in certain types of blocking oscillator circuits, adequate rise time of the triggering pulse is important for proper operation of the blocking oscillator.

In the two examples cited it is important to measure the input signal characteristics to determine whether a measurement on the assembly under test will be valid. Such an input test will be called a qualification test. The test on the assembly itself is called a performance test. In order to avoid an erroneous performance measurement on an assembly being tested, a qualification test at the input must be made simultaneously. Including additional comparators (cells) in the test instrument for making qualification measurements greatly increases the flexibility of FIST, since the operator is not bound by a testing sequence. To further increase the utility of the test instrument it should be possible to use any of the additional comparators as performance measurement cells so that simultaneous multiple performance measurements can be made.

## A FOUR-CELL TEST INSTRUMENT

This section describes a four-cell test instrument with explanations of programming, indications, and over-all operation.

Figure III-5 shows a simplified block diagram of a four-cell test instrument. The instrument has four cells of the type illustrated in Figure III-3 except that all cells share the same good - bad indicator. The instrument consists of four cells, three separate choppers for measurement of three separate voltages, a reference voltage, and a programmable indicator.

Four identical cells give the instrument a wide range of flexibility. The cell labeled master cell is always used in a performance test. The other three cells may be programmed individually into either performance or qualification cells. Often an assembly to be checked will require only the use of the master cell for a performance test and one other cell for a qualification measurement. On the other hand some assemblies may require the use of all four cells.

The programming of the four cells for either a performance or a qualification test is accomplished in a very simple manner. The master cell is always a performance cell while the other three cells (A,B, and C) may be programmed for use in either a performance or a qualification test, by the three programming connections (A,B,C,) of Figure III-5. Cells A,B, and C can be programmed as qualification cells by grounding their programming connections, or programmed into performance cells by connecting to their programming connections to the pin labeled P (performance). On the other hand if the three channels A,B, and C are not required, their programming connections are left floating. The interconnections among the programming pins are made by prewired jumpers

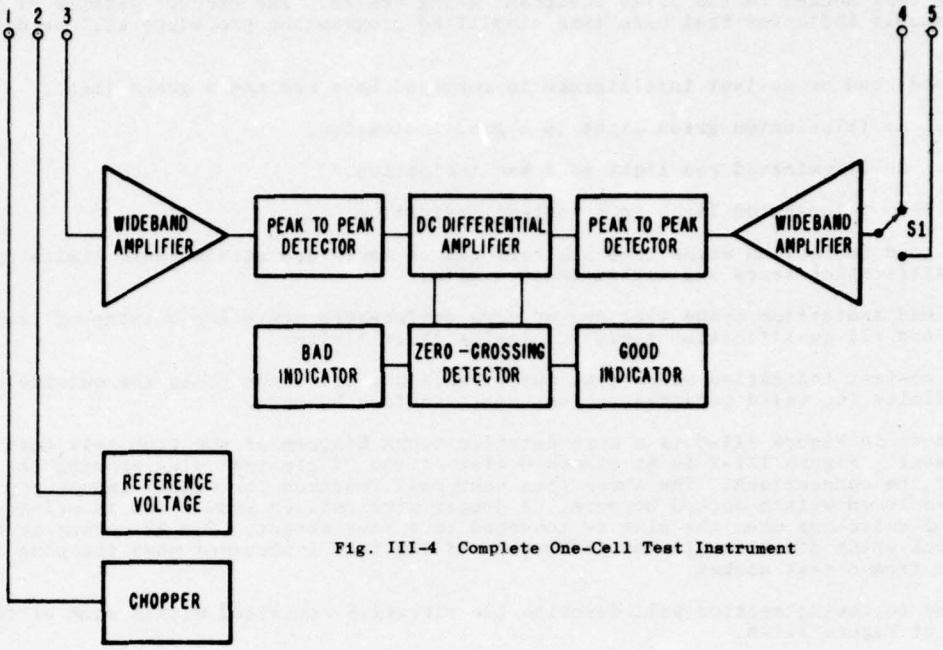


Fig. III-4 Complete One-Cell Test Instrument

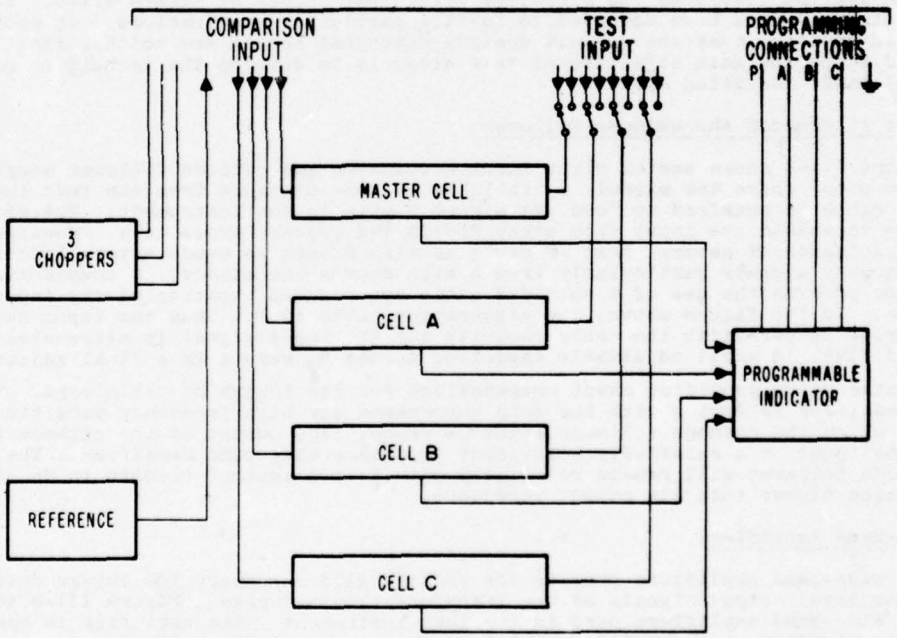


Fig. III-5 Block Diagram Of A Four-Cell Test Instrument

on each test socket in the prime equipment being tested. The circuit details of the programmable indicator that uses this simplified programming procedure will be described later.

Good, bad or no-test intelligence is conveyed by a red and a green light.

1. An illuminated green light is a good indication.

2. An illuminated red light is a bad indication.

3. No illuminated light is a no-test indication.

A good indication means that all performance tests are within their limits and all qualification tests are within their limits.

A bad indication means that one or more performance tests are outside of their limits and all qualification tests are inside their limits.

A no-test indication means that one or more qualification tests are outside of their limits (no valid performance test can therefore be made).

Shown in Figure III-6 is a more detailed block diagram of the four-cell test instrument. Figure III-7 is an expanded view of the 24 pin test plug showing the location of the connections. The three pins each cell requires for signal inputs are shown enclosed within dotted borders. A jumper wire between pins 3 and 14 offers a means of switching when the plug is inserted in a test socket. Pin 22 serves as an interlock which disables all the indicators of the test instrument when the plug is removed from a test socket.

The following section will describe the circuitry contained within each of the blocks of Figure III-6.

## CIRCUIT DESCRIPTIONS

This section describes the circuitry within each block of Figure III-6. Some special circuits have been designed to fulfill particular applications. It should be remembered however, that the circuit designs described herein are neither final nor optimized since the main objective of this study is to develop the techniques of a practical fault isolation system.

### 1. Input Attenuator and Cathode Follower:

Figure III-8 shows one of eight input attenuator and cathode follower stages. Since the point where the signal is available is some distance from the test instrument, shielded cable is required to feed the signal sample to the instrument. The shield is necessary to shield the input from stray fields and prevent cross talk. However, the shunt capacitance of several feet of cable is high enough to cause severe loading of high frequency signals particularly from a high source resistance. A compensated attenuator permits the use of a shielded cable and reduced capacity at the input to the probe. In the figure shown, the attenuation ratio is 5. Thus the input capacity at the probe is one-fifth the cable capacity and the input signal is attenuated by a factor of five. A small adjustable capacitor across $R_2$ serves as a final adjustment for the attenuator providing exact compensation for the length of cable used. The 100 ohm resistor in series with the grid suppresses any high frequency parsitic oscillations to which the cathode follower might be prone. The output of the cathode follower drives the input of a relatively high input impedance wide-band amplifier. The gain of the cathode follower will remain relatively gain stable against changes in Gm since its load is much higher than its output impedance.

### 2. Wide-Band Amplifier:

The wide-band amplifiers provide the voltage gain necessary for linear detection of the low level output signals of the transformation networks. Figure III-9 shows one of eight wide-band amplifiers used in the test instrument. The amplifier is comprised of three assemblies; a preamplifier which provides most of the voltage gain, a power amplifier which provides a large dynamic output and some additional gain, and a low impedance output driver.

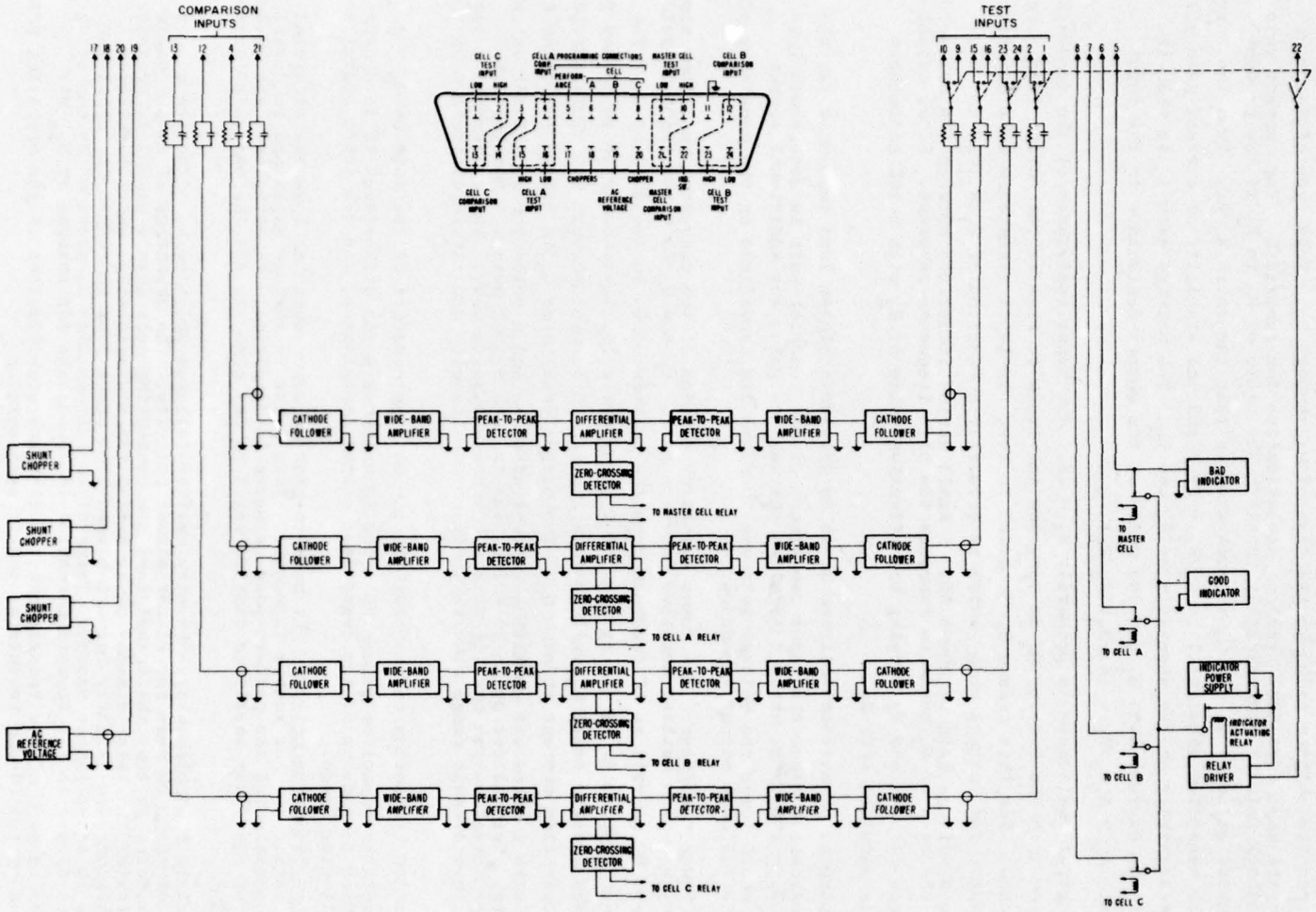


**Fig. III-6 Detailed Block Diagram of a Four-Cell Test Instrument**

(a) Preamplifier: The preamplifier consists of a feedback pair with input and output emitter followers $Q_1$ and $Q_4$. $Q_2$ and $Q_3$ form the feedback pair with inverse feedback from the emitter of $Q_3$ to the base of $Q_2$ through $R_2$. This type of feedback results in the most efficient trade of amplification for bandwidth. The current gain of the feedback pair is approximately equal to the ratio of $R_2$ to $R_3$ if the product of the current gains of $Q_2$ and $Q_3$ is much greater than the ratio $R_2/R_3$. Thus the current gain of the amplifier shown will be $R_2/R_3 = 12.5$ and the stability of current gain will depend substantially on the stability of $R_2$ and $R_3$. The voltage gain $A_v$ is then the product of the current gain $A_1$ and the ratio of the output resistance to the input resistance. $A_v = A_1 R_o/R_i = (R_2/R_3)(R_4/R_1)$.

The output resistance is primarily $R_4 = 1K$. The input resistance of the feedback pair looking into the base of $Q_2$ is very low and tends to rise as the cutoff frequency is approached. For this reason $R_1$ is added to keep the input resistance more constant. The expression for voltage gain, which is primarily a function of $R_1, R_2, R_3$, and $R_4$, results in a voltage gain of about 65. A small series inductance between the collector of $Q_2$ and the base of $Q_3$ tends to increase the high frequency response. Final adjustment of gain is obtained by varying the effective value of $R_3$ with a 10K adjustable resistor in parallel with $R_3$.

The purpose of emitter follower $Q_1$ is to provide a higher load impedance for the cathode follower. Since the input resistance of the feedback pair is determined primarily by $R_1 = 196$ ohms, which determines the voltage gain, any additional source impedance would lower the voltage gain and increase its dependence on the constancy of the cathode follower output impedance.

(b) Power Amplifier: The power amplifier portion of the wide-band amplifier also consists of a common-emitter negative feedback pair. $Q_5$ and $Q_6$ form the feedback pair with inverse DC as well as AC feedback from the emitter of $Q_6$ to the base of $Q_5$. The 10-volt regulator diode in the emitter of $Q_5$ provides a low impedance path to ground for the grounded emitter stage $Q_5$ while regulating the DC current through $R_7$. This provides a stable operating current through $Q_6$. The output transistor $Q_6$ is biased for class A operation with a quiescent operating current of 43 ma and a collector to emitter voltage of 40 volts. This allows up to 70 volts peak-to-peak output swings. However, the amplifier's normal output usually does not exceed 20 volts peak. The conservative use of the maximum dynamic range results in improved linearity and little distortion of signals.

The major limitation on bandwidth is the RC time constant of the load on $Q_6$. A small adjustable capacitor across R5 tends to compensate for differences in load capacitance, which is made up of the transistor output capacitance, and the input capacitance of the following stage.

(c) Low Impedance Driver: The peak-to-peak detector which follows the wide-band amplifier requires a low source impedance. This point is further explained in detail when the operation of the peak-to-peak detector is described. Nevertheless, the output impedance of the power amplifier itself, which is 820 ohms, is not low enough to drive the detector.

The class B complementary-symmetry emitter-follower provides an effective output impedance of about 20 ohms for either signal polarity. The advantage of using complementary symmetry is that the transistors are conducting only when a signal is present. This eliminates the large standby power needed for a single-stage emitter-follower. Both transistors are slightly forward-biased with about 2 ma to reduce any crossover distortion to a negligible amount. Temperature compensation is provided by operating two silicon diodes in the forward direction to compensate for changes of $V_{be}$ with temperature in each of the transistors. Additional stabilization of the operating point is achieved by two 10-ohm resistors, one in each emitter.

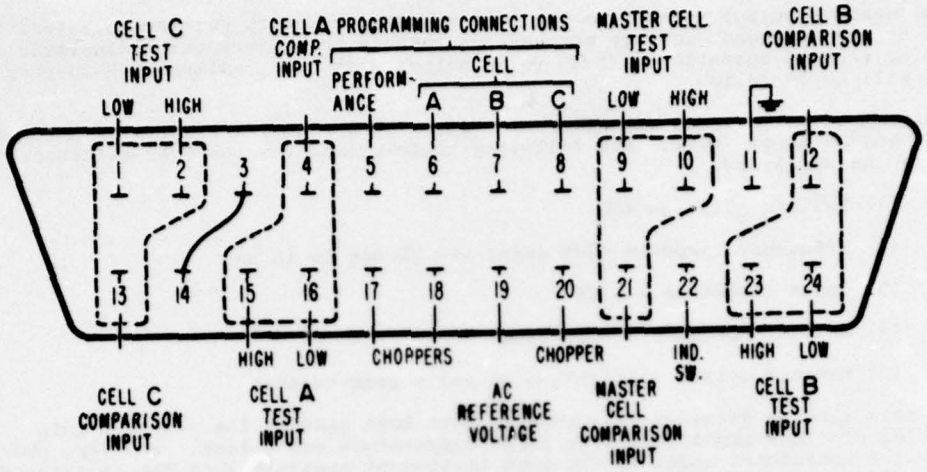


Fig. III-7 Expanded View Of Test Plug Pin Locations

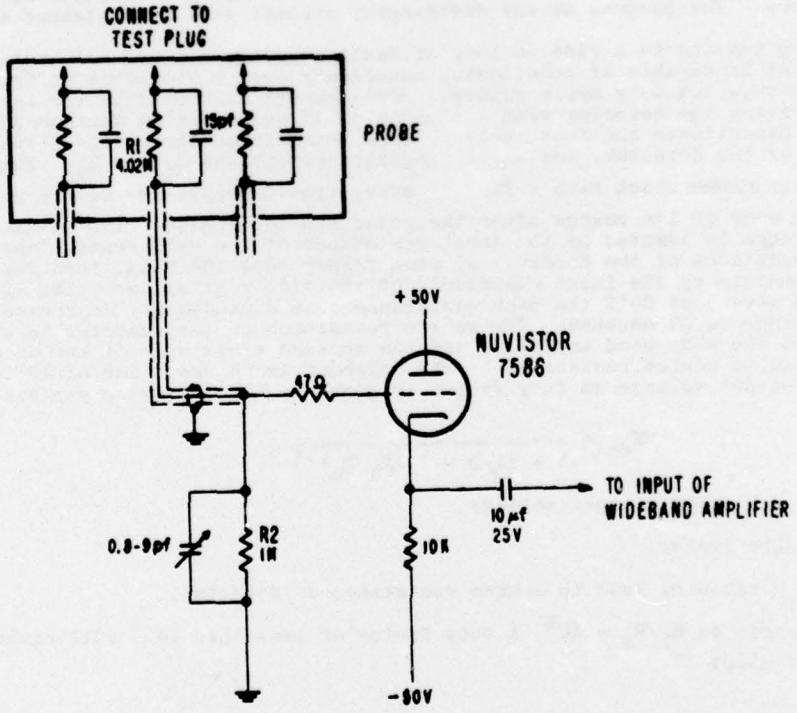


Fig. III-8 Input Attenuator And Cathode Follower

The maximum output swing from the complementary-symmetry emitter-followers is limited by the quiescent voltage across each transistor. Since each transistor is biased for class B operation with 25 volts across each, the maximum peak-to-peak output voltage will be 50 volts.

The complete wide-band amplifier of Figure III-9 requires 8 watts of power (88 ma at + 50v and 70 ma at -50v). The following table summarizes the over-all characteristics of the amplifier.

(1) Voltage gain --- 400

(2) Frequency response (3db down) --- 12 cps to 12 mc

(3) Input impedance --- 4K

(4) Output impedance --- 20 ohms

(5) Maximum output voltage --- 50 volts peak-to-peak

At this time no detailed measurements have been made on the relative gain stabilities of these amplifiers with large temperature variations. However, observations on the breadboard model of the test instrument indicate that the relative gains of the amplifiers remain very closely matched and quite satisfactory. The problem of maintaining matched relative gains of the two amplifiers is not as acute as stable absolute gain of a single amplifier.

3. Peak-to-peak Detector:

The peak-to-peak detector provides a positive DC output voltage which is proportional to the peak-to-peak amplitude of the signal from the wide-band amplifier. Figure III-10 shows the schematic daigram of the detector with a discharging circuit. There are eight such detectors (two for each cell) and one discharge circuit which supplies a pulse at the appropriate time to the discharging transistor across each of the detectors. The purpose of the discharging circuit will be explained subsequently.

For application in a wide variety of fault location measurement techniques, the detector must be capable of functioning accurately over a wide-band of frequencies and with very narrow, low duty cycle pulses. Best results are obtained for any given duty cycle by driving the detector with a minimum of 15 volts, using semiconductor diodes having low capacitance and fast recovery time, maximizing the ratio of load to source resistance of the detector, and maximizing the capacitance $C_1$ and $C_2$. The use of semiconductor diodes which have a fast recovery time prevents the memory capacitor $C_2$ from losing some of its charge after the pulse has terminated. The magnitude of the load resistance is limited by the input resistance of the differential amplifier and the back resistance of the diodes. At room temperature the total load resistance is determined mainly by the input resistance of the differential amplifier, which is 200 megohms. However, at 60°C the back resistance of the diodes may descrease the total load resistance to 50 megohms. The source resistance of the detector is the output impedance of the wide-band amplifier and the forward resistance of the diodes. The ratio of load to source resistance of the detector is in the order of $10^6$. The dependence of DC output voltage on duty factor is given by the following expression,[1]

$$E_{dc} = \frac{E_p}{1 + (1/D - 1)(R_1/R_s)^{-1}},$$

where $E_p$ = peak-to-peak input voltage,

$D$ = duty factor,

$R_1/R_s$ = ratio of load to source resistance of detector.

Thus for a ratio of $R_1/R_s = 10^6$, a duty factor of less than $10^{-4}$ will cause an error in DC output of about 1%.

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[1] Allan Easton, "Pulse Response of Diode Voltmeters," Electronics - January 1946.

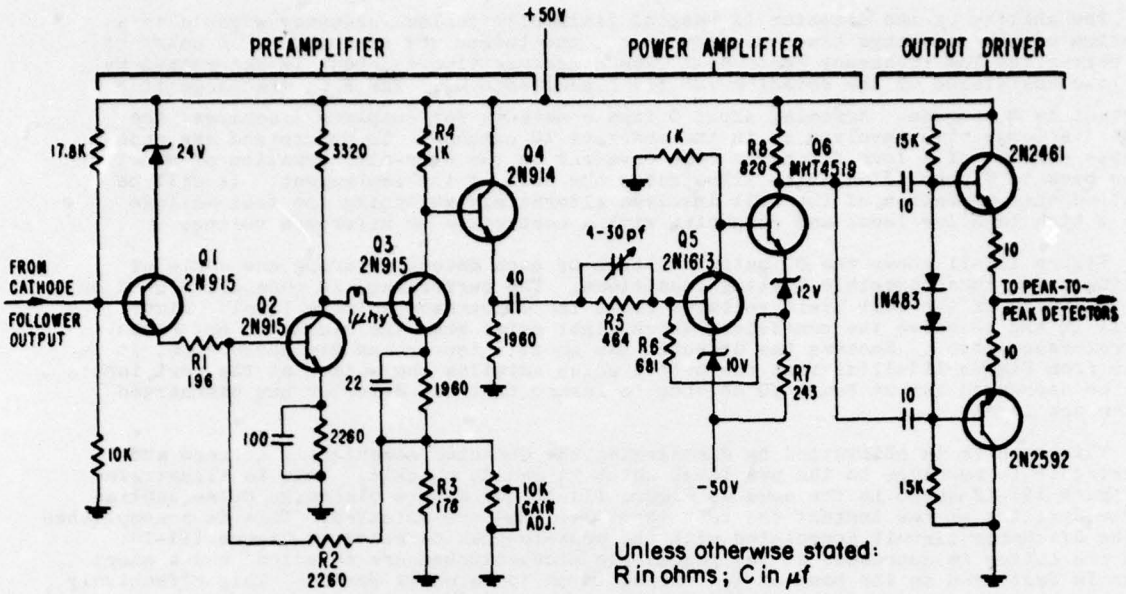



Fig. III-9 Wideband Amplifier

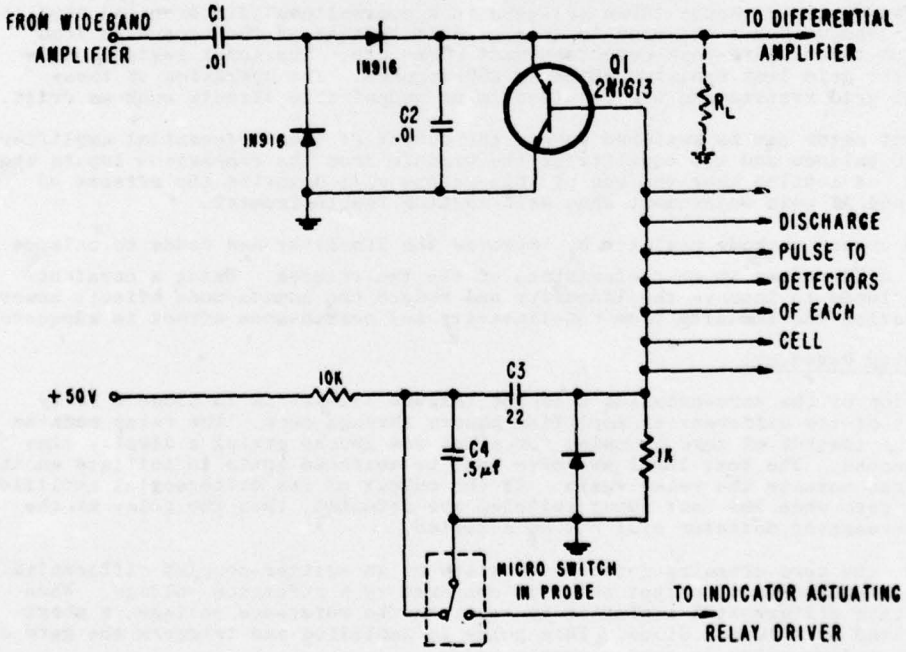



Fig. III-10 Peak-To-Peak Detector And Discharge Circuit

The ability of the detector to respond faithfully to low frequency signals is a function of the discharge time constant (i.e., the longer the discharge time constant, the better the low frequency response). The discharge time constant is determined by the load resistance of the detector and the capacitance $C_2$. The $R_L C_2$ discharge time constant is 2 seconds. Assuming about 5 time constants for complete discharge, the total discharge time involved is in the order of 10 seconds. To understand the undesirable effects of a long discharge time constant on the over-all operation of a cell, refer back to Figure III-4 which illustrates one cell of the instrument. It will be recalled that operation of the cell involves alternately switching the test voltage from a high to a low level and comparing with a comparison or reference voltage.

Figure III-11 shows the DC output voltage of each detector during one cycle of testing under three possible testing conditions. The performance is considered good or within limits if the test limit voltages cross the comparison voltage level. Figure III-11 (b and c) shows the conditions which might exist when the limits do not cross the reference level. Because the detector has about a ten-second discharge time, it is clear from Figure III-11(a) that the button which actuates the switch at the test inputs must be depressed for at least 10 seconds to assure that the detector has discharged to the new level.

This problem is alleviated by discharging the detector momentarily to zero and allowing it to recharge to the new level which it can do quickly. This is illustrated in Figure III-12 which is the same as Figure III-11 but with a discharge pulse applied to the detector at the instant the test input switches are actuated. This is accomplished by the discharge circuit associated with the peak-to-peak detector of Figure III-10. When the button is depressed at the probe, the microswitches are actuated, and a short pulse is delivered to the base of $Q_1$ turning it on for a short period. This effectively shorts the output of the detector to ground momentarily. After termination of the discharge pulse the detector quickly charges to the new level. The diode across the 1K resistor limits the destructive reverse voltage across the base to emitter junction of $Q_1$ to a safe level when the microswitch is returned to its normal position.

4. Differential Amplifier:

The differential amplifier (Figure III-13) provides an output signal which is proportional to the difference between the signals from the detectors that appear at each input. Two Nuvistor vacuum tubes are used in a conventional differential amplifier configuration. Vacuum tubes are used in this circuit because of the very high load resistance which the peak-to-peak detectors must work into. The input resistance is determined by the grid leak resistor which is 200 megohms. The operation of these tubes with high grid resistances has resulted in no undesirable effects such as drift.

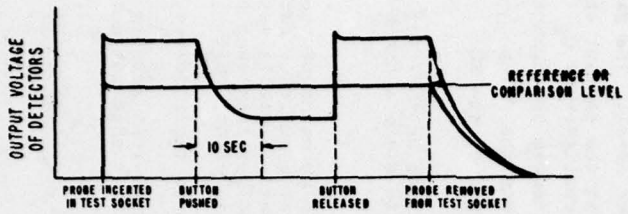
The current meter can be switched across the output of the differential amplifier to check the DC balance and the equality of the signals from the comparison inputs and the test input. A section near the end of this report will describe the effects of zero balance, and AC gain adjustment when self-testing the instrument.

The large common-cathode resistor $R_1$ improves the linearity and tends to balance the effects of differences in characteristics of the two triodes. Using a constant-current source tends to improve the linearity and reduce the common-mode effect; however, in this application the immunity from non-linearity and common-mode effect is adequate.

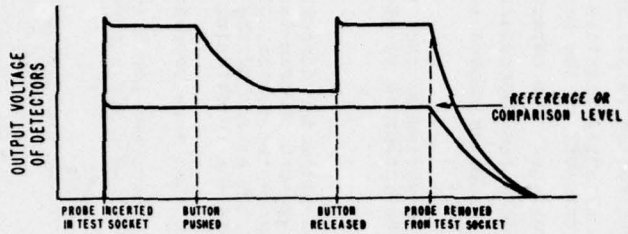
5. Zero-Crossing Detector:

The function of the zero-crossing detector (Figure III-14) is to close a relay when the output of the differential amplifier passes through zero. The relay remains closed after the instant of zero crossing for about one second giving a display time of about one second. The test input switches must be switched again to initiate another zero crossing and actuate the relay again. If the output of the differential amplifier does not cross zero when the test input switches are actuated, then the relay at the output of the crossover detector will not be actuated.

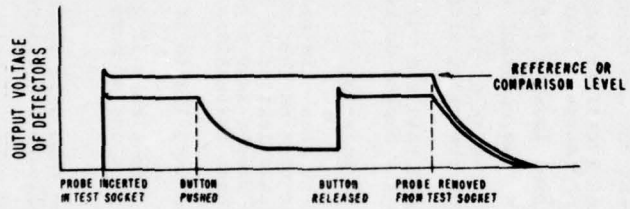
Basically, the zero-crossing detector consists of an emitter-coupled differential amplifier with a single-ended output which is compared to a reference voltage. When the output of this differential amplifier is equal to the reference voltage, a short pulse is generated by a tunnel diode. This pulse is amplified and triggers the gate of a controlled rectifier which in turn closes a relay.




(a) GOOD PERFORMANCE LIMITS CROSS REFERENCE LEVEL

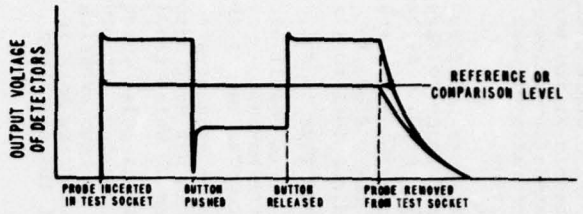



(b) BAD PERFORMANCE LIMITS DO NOT CROSS REFERENCE LEVEL

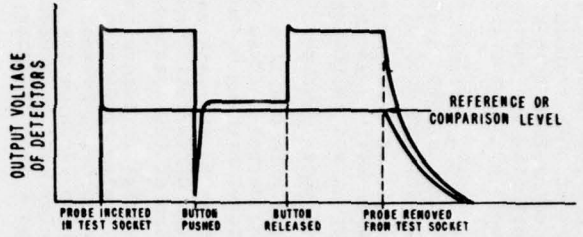



(c) BAD PERFORMANCE LIMITS DO NOT CROSS REFERENCE LEVEL

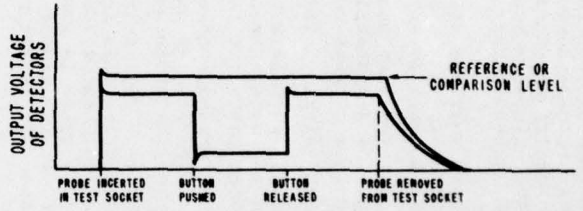
Fig. III-11 Detector Response to Three Possible Testing Conditions with Normal Discharging




(a) GOOD PERFORMANCE LIMITS CROSS REFERENCE LEVEL




(b) BAD PERFORMANCE LIMITS DO NOT CROSS REFERENCE LEVEL




(c) BAD PERFORMANCE LIMITS DO NOT CROSS REFERENCE LEVEL

Fig. III-12 Detector Response to Three Possible Testing Conditions with Detector Rapid Discharging Employed

The differential amplifier uses two high-gain, silicon, planar transistors in one enclosure. This type of construction offers good stability against temperature changes, since the temperature dependent parameters of both transistors are nearly at the same temperature. An additional advantage of using this type of dual transistor enclosure is due to the manufacturer's success in matching the current transfer ratio of each transistor. The 2N2060 is particularly well suited for this application.

When the input voltages to each base of the differential amplifier transistors are equal, an output voltage will exist which is just equal to the reference voltage determined by the 25-volt zener diode, the forward drop of $CR_1$ and $V_{be}$ of $Q_3$. As base 1 becomes more positive than base 2, a result of crossing zero in one direction, the output voltage of the differential amplifier will cause the 25-volt zener diode to conduct, causing base current to flow into $Q_3$ and switching tunnel diode TD-1 to its forward conduction voltage. This positive pulse reverse biases $Q_3$ and thus is not amplified. However, as base 2 then becomes more positive than base 1, (a result of crossing zero in the other direction), the output voltage of the differential amplifier will drop below the reference level causing $Q_3$ to turn off. The tunnel diode switches from the forward conduction voltage to the peak voltage, which generates a negative pulse. However, the output of the differential amplifier can not drop below the reference level instantaneously because of the feedback capacitor $C_1$ which tends to delay this change. This delay is essential to insure that the narrow pulse which discharges the detectors does not cause a false indication of zero crossing. In other words, there must be about a 200 millisecond delay between the time of zero crossing and the time when a pulse is generated which turns on the relay. Referring back to Figure III-12(b), it is evident that a delay must be present so that the discharge pulse from the detector is not recorded as a zero crossing.

After suitable delay, the negative pulse generated by the tunnel diode is amplified and inverted by transistor $Q_4$. The output pulse from $Q_4$ triggers the controlled rectifier into conduction thereby discharging $C_2$. The 150 ohm resistor in series with $C_2$ limits the peak current through the controlled rectifier to a safe level. The gate of the controlled rectifier returns to a slightly negative voltage at the instant the trigger pulse terminates. This insures turn-off of the controlled rectifier as $C_2$ begins to charge up towards the supply voltage. The capacitor $C_2$ charges up through the relay with a time constant determined by the capacitance of $C_2$ and the relay resistance. The length of time the relay remains closed is determined by the RC charging time and the relay drop-out current.

The design of this circuit is being modified so that the operation of the test instrument does not depend on relays. The intent is to use all solid state switches.

6. Programmable Indicator:

Figure III-15 shows the programmable indicator circuitry, which by means of the four programming pins on the test set plug, can program the output relay of each cell into either a qualification or a performance indication. The indicator logic is set up to light the proper lamp depending on the action of the output relays from each cell. The following three indications illustrate the functioning of the indicator logic:

a. A green lamp lights if all cells that were programmed actuate their relays.

b. A red lamp lights if all cells programmed for a qualification test actuate their relays and if one or more cells programmed for a performance test does not actuate its relay.

c. Neither lamp will light if any of the cells programmed for a qualification test does not actuate its relays.

To understand the circuit operation, we first will assume that the programming connections 5,6,7,8, are floating. When the button is pushed the indicator actuating relay (shown in dotted lines) is closed providing a path for current to flow into the base of $Q_1$ and $Q_3$. However, the current flowing into the base of $Q_1$ will be much less

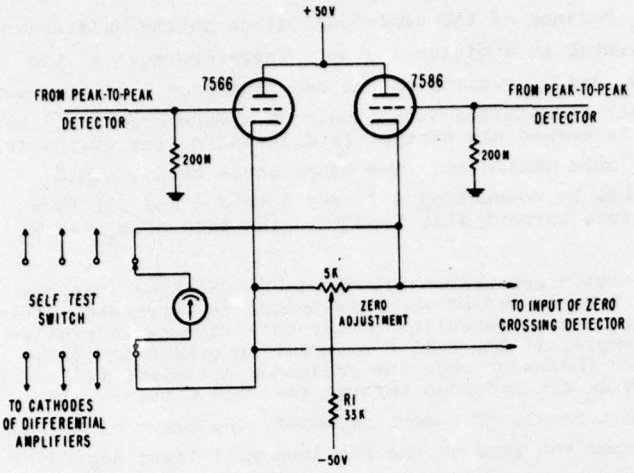



Fig. III-13 DC Differential Amplifier

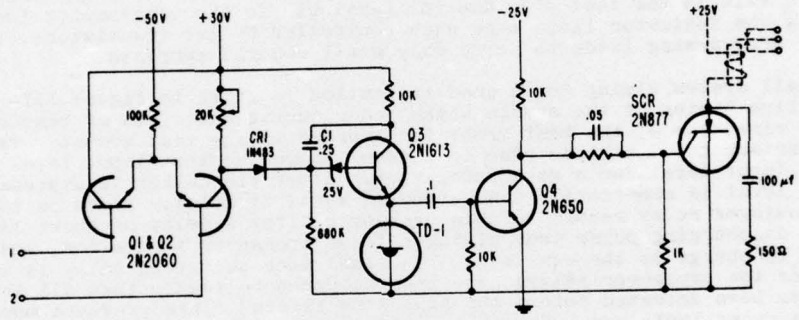



Fig. III-14 Zero Crossing Detector

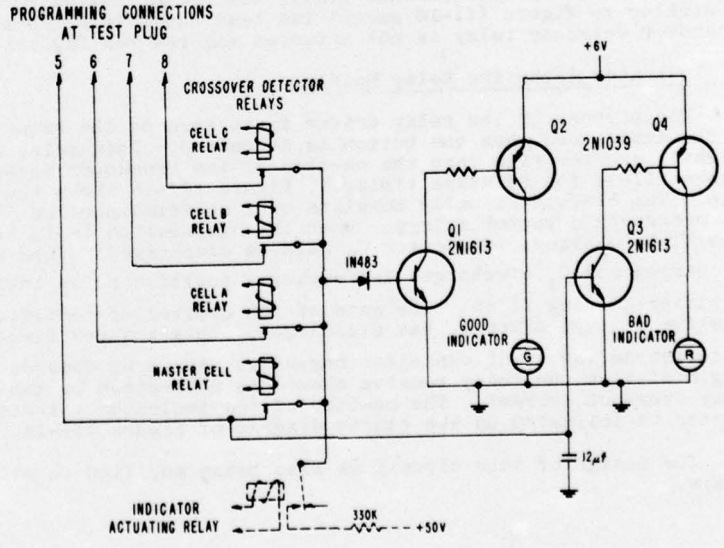



Fig. III-15 Programmable Indicator

than the current into the base of $Q_3$ because of the additional diode in the emitter-to-base connection between $Q_1$ and $Q_2$ causing an additional drop. Therefore most of the current will flow into the base of $Q_3$ and $Q_4$ turning on the bad indicator. If, however, the crossover relay of the master cell is actuated as a result of a zero crossing, the connection to the base of $Q_3$ and $Q_4$ is opened and current is allowed to flow only into the base of $Q_1$ and $Q_2$ turning on the good indicator. The other three cells, A,B,C, may be programmed for performance tests by connecting 6,7, and 8 to 5. Now all four crossover relays must be actuated before current will flow into the base of $Q_1$ and $Q_2$ to turn on the good indicator.

To program a cell for a qualification measurement it should be recalled that the indication requires that neither light be turned on when the button is depressed. This is accomplished by grounding the programming connection of any cell that is to provide a qualification measurement. For example, if the cell C programming connection 8 is grounded, current cannot flow to either indicator when the indicator actuating relay turns on, because the bases of $Q_1$ and $Q_3$ are grounded through the cell C relay. If, however, the cell C relay operates as a result of a zero crossing, the bases of $Q_1$ and $Q_3$ will no longer be grounded and either the good or the bad lamp will light depending on whether or not the master cell relay is also actuated.

It is found that it was necessary to keep the power low in the programming leads to reduce cross talk in the test plug due to transients in the programming leads. To accomplish this the indicator lamps were each controlled by two transistors, thereby permitting the programming leads to carry only small control currents.

The over-all system timing for a good indication is shown in Figure III-16. This shows the relative timing of the events which occur during one cycle of testing. At some arbitrary time, t = 0, the test probe is inserted in the test socket. The test input and comparison input signals rise to a level where the test input level is above the comparison input level for a good indication. After the button is depressed and the test input level is momentarily discharged to zero, it quickly rises to the new level. The crossover relay reacts to this crossover after a delay of about 140 msec. to prevent the discharging pulse from giving a false crossover indication, and to permit the test input to charge to the new level. The indicator actuating relay is turned on shortly after the crossover relays have been actuated to insure that all the crossover relays have been actuated before the indicator lights. This prevents momentary flashing of the wrong indicator. For the same reason the indicator is extinguished slightly before the crossover detector relays drop out. The system timing figure also shows that the button must be held in the depressed position only until the indicator lamp lights. In practice this requires only a momentary depression of the test button to get an indication. If the button is continuously depressed and released before the indicator normally extinguishes itself the proper indicator will remain on. Figure III-17 is similar to Figure III-16 except the test input level does not cross zero; hence the crossover detector relay is not actuated and the red light will light.

7. Indicator Actuating Relay Driver:

The purpose of the relay driver is to turn on the relay which controls the power to the indicators when the button is depressed. This relay remains latched for a slightly shorter time than the on-time of the crossover detector relays. (Refer to Figure III-14 for relative timing.) Figure III-18 shows the circuit which accomplishes this. The circuit actually consists of a modified portion of the circuit used to drive the crossover detector relays. When the microswitch is in its normal position, capacitor $C_1$ will be charged. Capacitor $C_2$ remains discharged. When the microswitch is actuated, $C_2$ charges and $C_1$ discharges delivering a positive pulse into the gate of the controlled rectifier turning it on. The gate of the controlled rectifier returns to a slightly negative voltage after $C_1$ has discharged. This insures turn-off of the controlled rectifier as the 75 $\mu$f capacitor begins to charge up towards the supply voltage. The length of time the relay remains closed is determined by the RC charging time and the relay drop-out current. The on-time of the indicator actuating relay is somewhat shorter as indicated on the timing diagram of Figure III-16.

The design of this circuit is also being modified to eliminate electro-mechanical relays.

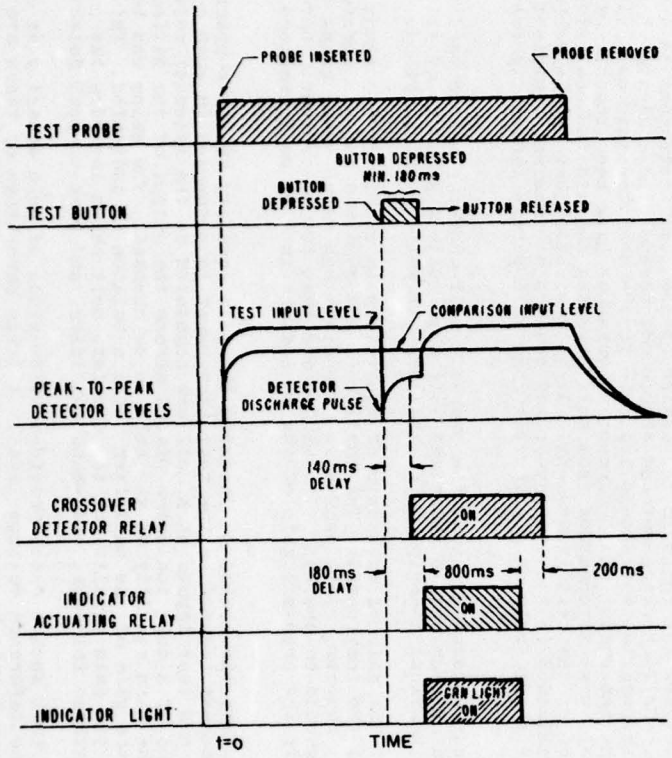



Fig. III-16 Instrument Timing Sequence For A Good Indication

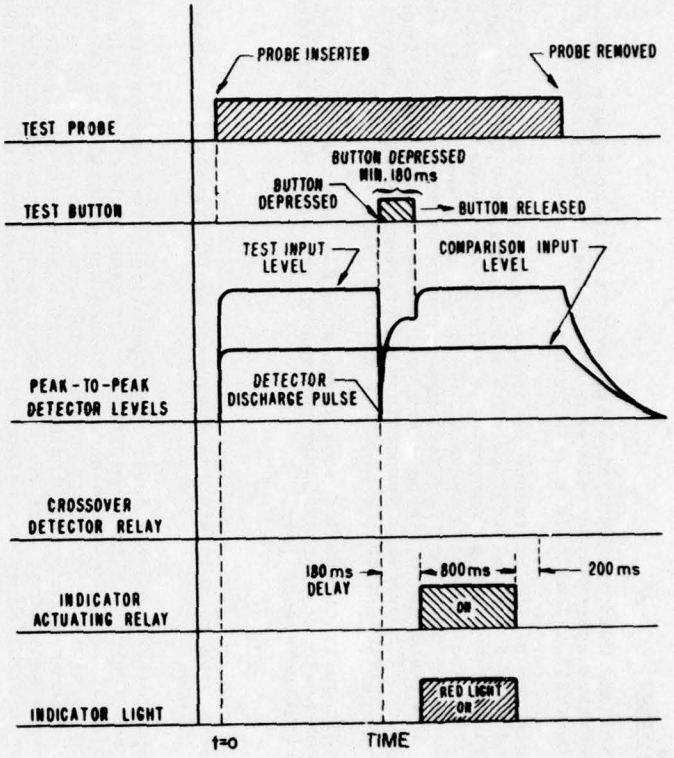



Fig. III-17 Instrument Timing Sequence For A Bad Indication

8. AC Reference Voltage:

The reference voltage used in the test instrument is a 10 kc square wave voltage derived from a temperature compensated reference diode which is converted to a square wave by means of a conventional astable multivibrator.

The schematic diagram of the reference generator is shown in Figure III-19. The 1N3155 reference diode has a nominal voltage of 8.4 volts with a 0.005%/C° temperature coefficient. The transistors used to convert the reference voltage to a square wave have a finite saturation voltage which is about 5% of the reference voltage. This is a moderate potential source of error; however, measurements on the output of the reference generator show about a 0.02%/C° temperature coefficient of the whole circuit which includes the attenuator. The over-all temperature coefficient as given above is sufficiently low and compatible with the accuracy goals of this test set.

The reference voltage is attenuated to a level of 0.240 volts which is the preferred input level to the comparators. The variable portion of the attenuator provides fine adjustment of the output reference level.

9. Shunt Choppers:

The shunt choppers used to modulate low frequency or DC signals for measurement are shown in Figure III-20.

Solid state choppers are used for their long operating life compared with electromechanical choppers and their ability to operate at the same frequency as the AC reference generator which is 10 kc. The three shunt choppers are individually driven by separate transformers to minimize coupling between choppers. The 10 kc driving voltage is obtained from the reference generator through a buffer amplifier $Q_4$ to minimize the loading on the reference generator. The choppers must be operated within their current ratings. This can be controlled by the networks in the transformation networks.

## SELF-CHECKING THE TEST INSTRUMENT

One of the objectives set forth in the design of the test instrument is a self-checking capability. Self-checking methods should not necessarily indicate what portion of the test set is inoperative but should indicate to the operator that no further use should be made of the instrument until corrective measures have been taken. Of course, where it is practical, some information from self-checking can aid in localizing trouble to a specific portion of the test instrument. It is desirable that the self-checking circuitry be self-contained in the instrument and require no external equipment. The self-checking method should rely on a principle that will permit inherently reliable testing.

To obtain a clear understanding of the over-all self-checking method used in this system, a discussion of each single important self-check with a sequence of illustrations will precede the final description of the over-all self-checking method.

The equality of the gain of each pair of amplifiers in a cell is important to the over-all accuracy of the instrument. The amplifiers are designed for a moderately high gain stability; however, small drifts in components can cause their gains to drift apart. This should be checked. It was decided that the system would have adequate stability if the over-all gain of the amplifiers in each cell remained within 1% of each other.

Figure III-21 shows the method of checking the gain equality of one comparator cell. A self-test socket located on the test instrument is connected in such a manner that the comparison and test inputs of a cell are connected to the internal reference voltage. Observation of a null indicator placed across the output of the differential amplifier permits the gain equality of the cell to be checked. The gains can be made equal by adjusting the gain of one amplifier until a balance is indicated. This test provides a check on the gain equality of the complete cell which includes the compensated attenuator, cathode follower, wide-band amplifier, and peak-to-peak detector.

It is realized that such a check provides information of gain equality at one input level only, the reference voltage level. A large percentage of tests are made at the reference input level; however, some comparisons at other levels are also frequently made. The check on gain equality at one level assumes reasonable good linearity. However, again, the major concern is not the absolute linearity since two identical

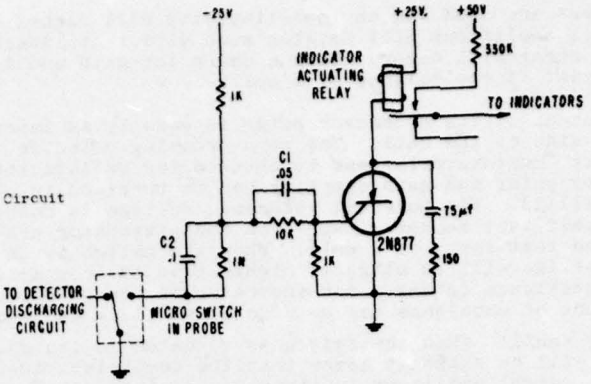



Fig. III-18 Indicator Actuating Relay Driver Circuit

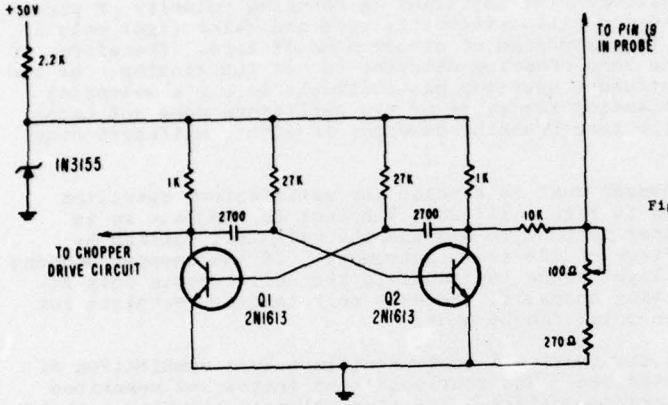



Fig. III-19 Reference Voltage Generator

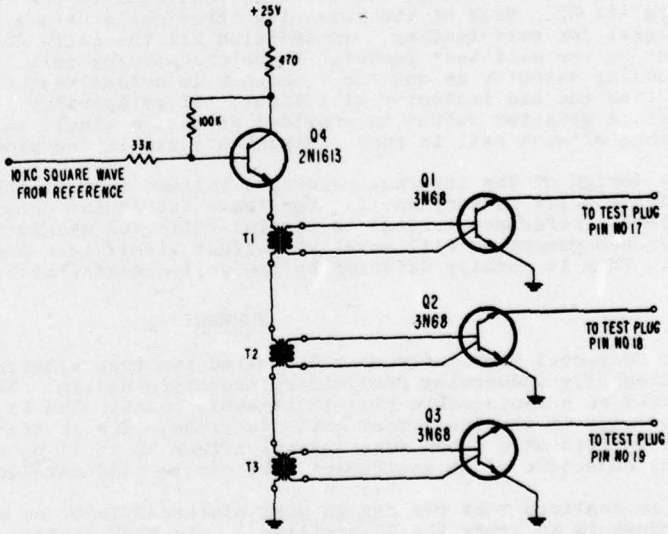



Fig. III-20 Shunt Chopper Circuitry

amplifiers are used and any non-linearity will cancel out. It is unlikely that two identical amplifiers will develop such different linearity characteristics that a self-testing error will occur. Thus a check for gain equality at one level should provide a good test of the balance of a cell.

A check of the crossover point is equally as important as a check of gain equality of each side of the cell. The zero-crossing detector, although designed with an inherent stability, must nevertheless be checked for satisfactory operation. A check of the crossover point and gain equality can be combined in one test as may be seen by examining Figure III-22. The internal reference voltage is connected to an attenuator located on the self-test socket. Taps from the attenuator are connected to the comparison input and test input of a cell. When the switch is in the position shown, the test input voltage will be slightly higher than the comparison input voltage causing a slight unbalance in the meter indicator at the output of the differential amplifier. The amount of unbalance can be made as small as desired by making $R_1$ and $R_2$ identical and very small. When the switch is actuated to its second position, the test input voltage will be slightly lower than the comparison input, causing the indicator to assume a slight unbalance in the other direction. Thus if the amplifiers are balanced, the indicator meter pointer movement each side of zero will be equal as the switch is actuated. When the output of the differential amplifier is changing polarity or passing through zero, the zero-crossing detector will actuate the good indicator light only if the zero-crossover point is within the excursion on either side of zero. Therefore, if the amplifiers are unbalanced or the zero-crossing detector is not functioning, the bad indicator will light. The test instrument operator has available to him a balancing control to balance the cell. If balancing the gains of the amplifiers does not cause the good indicator to light, then the zero-crossing detector is either malfunctioning or misadjusted.

The choppers in the test instrument must be checked for satisfactory operation also. This is accomplished as shown in Figure III-23. The test is the same as in Figure III-22, except that the chopper is used to provide the self-test voltage by modulating an attenuated supply voltage of the test instrument. If the chopper becomes inoperative the bad indicator will light since the input to the cell accepts only AC signals. When the chopper is operating normally, the same self-testing technique for amplifier balance and zero-crossover point can be made.

The self-checking technique on the four-cell instrument is only a combination of the self-testing techniques just described. The four-cell test instrument described contains three choppers and one reference voltage. The over-all self-checking technique used to test the whole instrument is accomplished with one self-test socket located on the test instrument. One of the cells is connected to the reference voltage as shown in Figure III-22. Each of the remaining three cells uses a chopper to provide the input signal for self-testing. In addition all the cells are programmed for a performance test on the self-test socket. Therefore, if any cell is out of balance, or the zero-crossing detector is out, or a chopper is defective, or the reference voltage is absent, then the bad indicator will light. If an operator gets a bad indication on self-test, a selector switch is provided so that a single internal meter can check the balance of each cell in turn. Separate controls are provided to balance each cell.

The design of the internal reference voltage is such that the probability of reference voltage drift is very small. For those infrequent occasions when one may wish to calibrate the reference voltage, a digital voltmeter should be used. Degradation of the reference generator will usually manifest itself as a complete loss of reference voltage. This is readily detected by the self-checking method previously described.

## SUMMARY

The four-cell test instrument described has been constructed in breadboard fashion and is presently undergoing preliminary prototype design. Each of the circuits is constructed on a replaceable plug-in assembly board. The Figure III-24 shows the projected size of the instrument with its probe. The instrument will occupy approximately one-fifth of a cubic foot and weigh from 12 to 14 pounds. The instrument easily meets the objective of an instrument that can be hand-carried.

It is realized that the design just discussed is by no means optimum since the main purpose is to study the feasibility of the FIST system. Certain improvements and modifications are currently being studies and applied. Some of the improvements which are under development are: automatic switching at the probe to eliminate the hand-actuated switch, elimination of electro-mechanical relays, and circuit redesign with

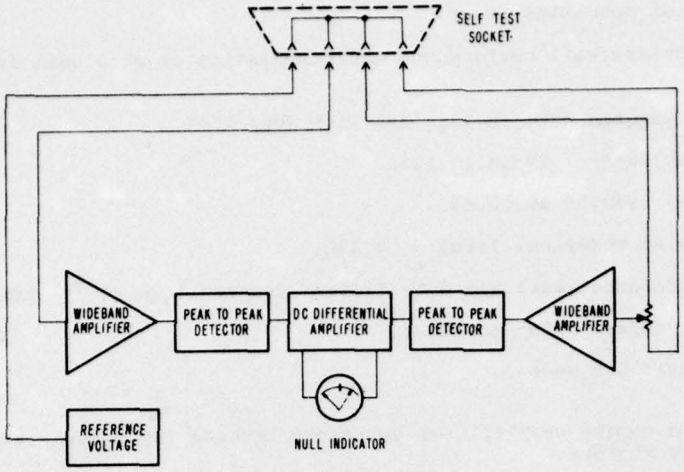



Fig. III-21 Self Testing For Gain Equality

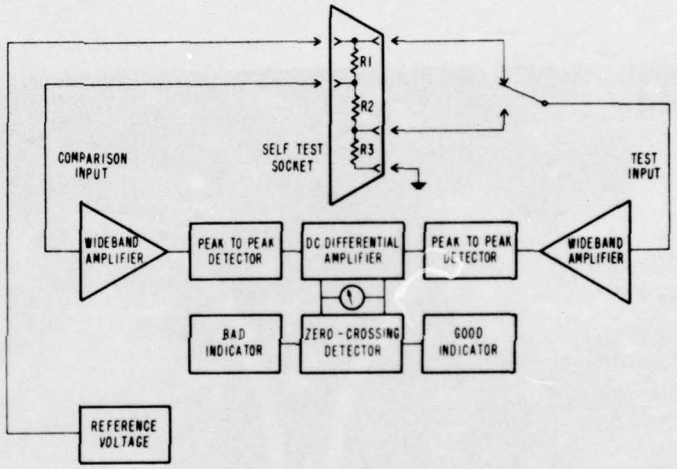



Fig. III-22 Self Testing Zero Crossing Point and Gain Equality

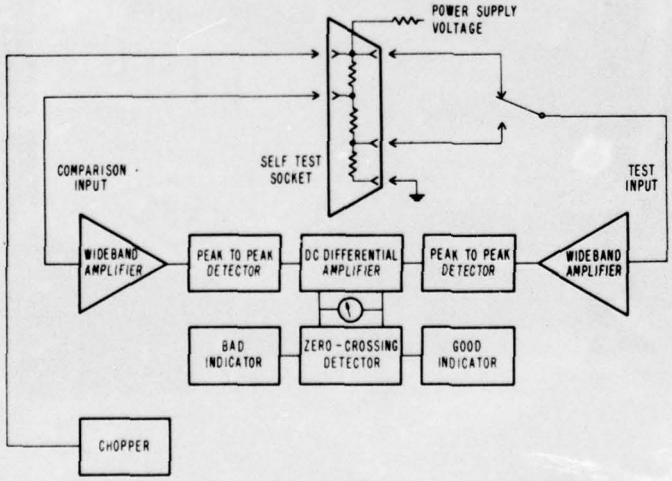



Fig. III-23 Self Testing Chopper, Zero Crossing Point, and Gain Equality

goals of simplicity and economy of operating power.

The following summarizes the over-all performance characteristics of each cell in the test instrument:

1. Input impedance: 5.02 megohms ±2%, in parallel with 20pf %10%.
2. Over-all band-width (3db down): 12cps to 12mc.
3. Reference voltage level: 240.00 mv ±0.4%.
4. *Precision of comparison at reference* level: < 0.25%.
5. Absolute accuracy at reference level and duty factors greater than $10^{-3}$: ±2%.
6. Useful input operating range:* 30mv to 300mv.
7. Minimum measurement time: 0.5 seconds.

* At lower input levels the drive to the peak-to-peak detectors becomes low enough that accuracy may suffer at low duty factors.

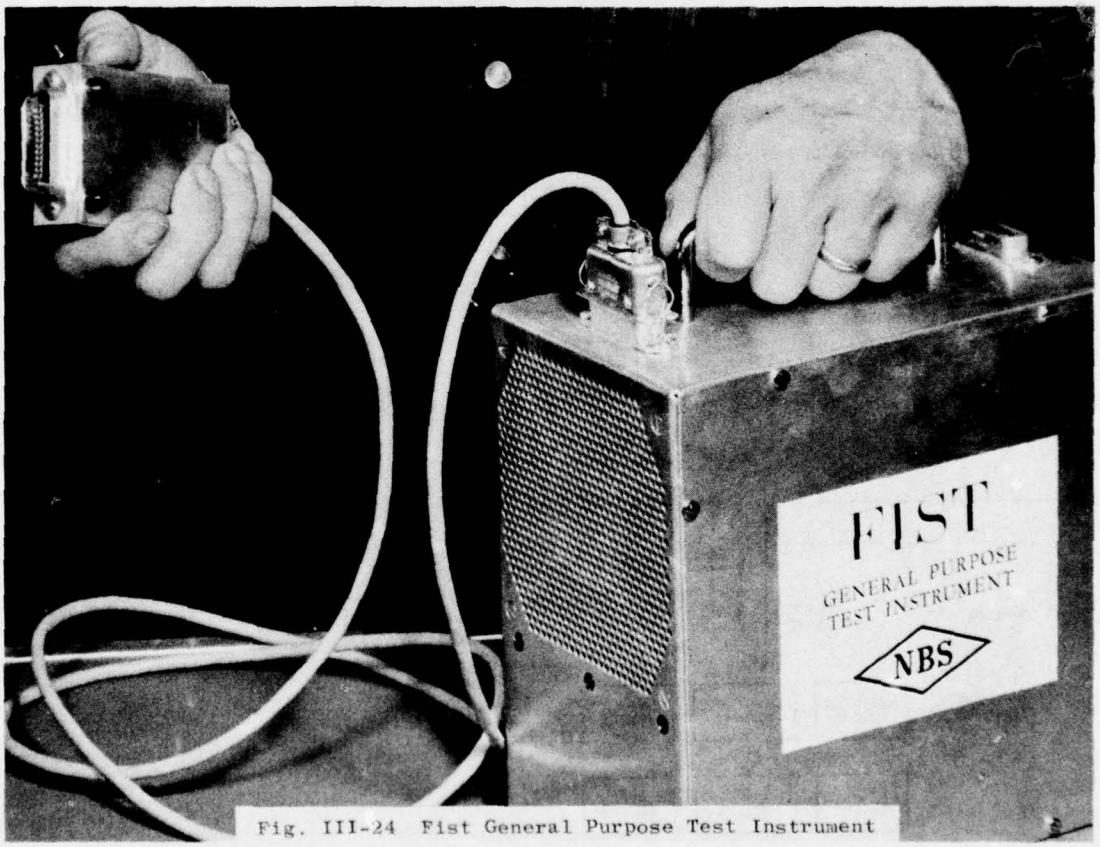



Fig. III-24 Fist General Purpose Test Instrument